RADIO
Handbook

EIGHTH EDITION

15¢

THE

"RADIO"

HANDBOOK

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By the Editors of "Radio"

"RADIO"

"RADIO" AMATEUR NEWCOMER'S HANDBOOK

"RADIO" NOISE REDUCTION HANDBOOK

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THE "RADIO" HANDBOOK

EIGHTH EDITION

ISSUED ANNUALLY

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THE "RADIO" HANDBOOK

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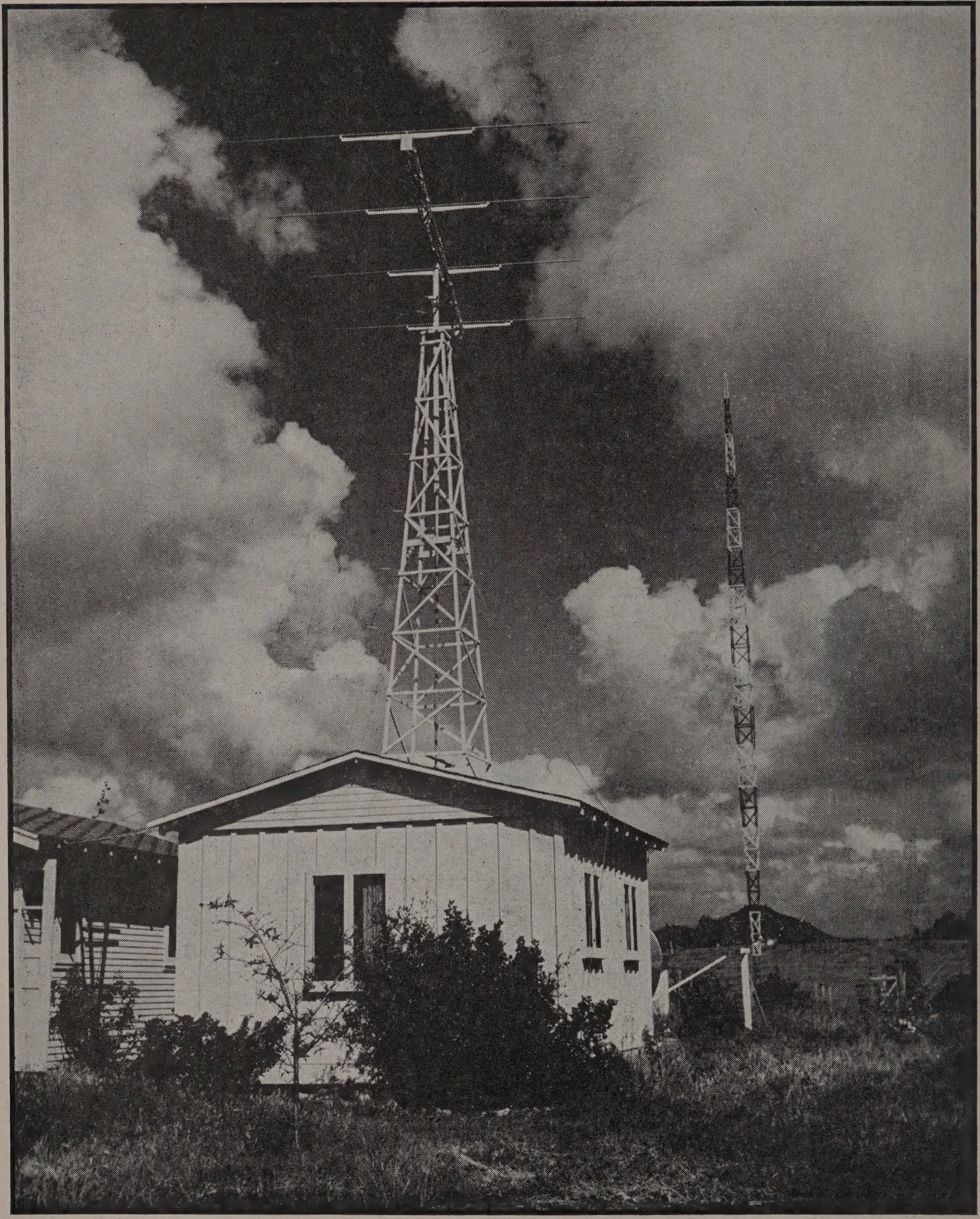
WRITTEN BY THE EDITORS OF "RADIO"



The amateur "shack" has become a familiar part of the American landscape. That of W6JRM is typical of many hundreds throughout the country.

CHAPTER ONE

Introduction to Amateur Radio

WHILE the definition of "amateur" would seem to include shortwave listeners as radio amateurs, the term ordinarily is used to indicate specifically those radio hobbyists possessing a government license and amateur call letters.

More than 50,000 licensed amateurs in the U.S.A. are actively engaged in this field for purposes of experimentation, adventure, and technical education. It is interesting to consider what there is about amateur radio that captures and holds the interest of so many people throughout the world and from all walks of life, for unquestionably there is something about it which generates a lasting interest in its varied problems and activities.

Many famous men, holding high-salaried positions of importance in the radio industry today, got their start in the radio business by discovering an interest in amateur radio. A large number of these executives and engineers continue to enjoy amateur radio as an avocation even though commercially engaged in the radio industry, so strong is the fascination afforded by this hobby.

Technical Achievement. Although "hamming" generally is considered to be "only a hobby" by the general public, its history contains countless incidents of technical achievements by its members which have served to improve radio communication and broadcasting. Many of the more important advancements in the art of radio communication can be chalked up to the ingenuity of radio amateurs. Experiments conducted by inquisitive amateurs have led to important developments in the fields of electronics, television, radio therapy, sound pictures, and public address, as well as in radio communication and broadcasting.

Fellowship. Amateurs are a most hospitable and fraternal lot. Their common interest makes them "brothers under the skin" and binds them together as closely as would membership in any college fraternity, lodge, or club. When visiting a strange town an amateur naturally first will look up any friends in that town he has made over the air. But even if he is unknown to any amateurs in that town, his amateur call is an "open sesame." The local amateurs will hang out the welcome sign and greet him like a long lost brother.

It is not unusual for an amateur to boast a large circle of friends, scattered throughout the country, with whom he chats nightly while seated comfortably at home. He gets to know these people intimately, many of whom he will never meet personally. Frequently he is of service to them, and they to him, in delivering messages to other people.

Amateur radio clubs have been formed in nearly all of the principal cities in the United States. The first thing a newcomer should do is to attend one of these club meetings and let the members know that he is interested in joining the ranks of radio amateurs. The veteran amateurs will be glad to lend a hand with any difficult problems he might encounter, and often can give invaluable advice as a result of their own experience. Also, he will be introduced to others who have recently taken an interest in amateur radio, and will have someone with whom to study. A "study companion" is especially helpful when it comes to learning the code.

The National Youth Administration is active in encouraging and assisting young men to become radio amateurs, and it is suggested that young, prospective amateurs in the larger cities contact the local N.Y.A. organization.

Public Service. The radio amateur, or "ham," often renders public service. When hurricane, flood, earthquake, or heavy ice wreaks havoc with telephone and telegraph lines and the mails, the newspapers invariably follow with an account of how aid was summoned to the devastated area and communication maintained with the outside world largely through the efforts of radio amateurs. Radio amateurs are justly proud of their record of heroism and service in times of emergency. Many expeditions to remote places have kept in touch with home and business by "working" amateurs on the short waves.

A Diversified Hobby. Amateur radio is a hobby with several phases. There are those who revel in long-distance contacts with amateurs in far-off lands and try to excel in number of distant stations "worked." These enthusiasts are called "dx" men. Unfortunately this activity has been curtailed by the war.

Others make a specialty of relaying messages free of charge for people in their communities, and these fellows often perform meritorious services. Still others prefer not to specialize, but simply to "chew the rag" with any other hams who happen to be on the air.

Then, there are the experimenters, indefatigable individuals always striving for perfection. They are everlastingly building up and tearing down transmitters and receivers, deriving as much enjoyment from the construction or improvement of equipment as from its operation on the air. Whichever phase most strongly captures your fancy, you will find amateur radio an absorbing hobby.

Before you may join the others on the air, however, you must be licensed by the government to operate a transmitting station; so your first task will be to acquire sufficient knowledge to pass the test. Those who attempt to operate (on the air) *any kind of transmitting* equipment without a license are liable to a fine and imprisonment.

How to Obtain Your License

To obtain an amateur transmitting license from the U.S. government, you must be a citizen of the U.S.A., master the code, know how amateur transmitters and receivers work and how they must be adjusted, and be familiar with regulations pertaining to amateur operators and stations. An application blank for amateur radio operator and station license can be obtained from your district office of the Federal Communications Commission. A list of district offices is printed in Chapter 29.

When you have filled out this application properly, sworn to it before a notary public, and returned it to the district office, the inspector in charge will notify you of the time and place of your examination. There is no charge for an amateur operator and station license, and there are no age limits.

It is necessary that your station not be located on premises under the *control* of an alien. Remember this when determining the proposed site of your transmitter and filling out the application blanks. If you rent from an alien, the premises are under your "control" and you have nothing to worry about. However, if you merely "board" instead of rent, that does not put the premises under your control.

The examination will consist of a practical code test and a written theoretical examination. The written examination usually includes ten questions, some of which are in several parts. The questions are of the "multiple answer" type, and the applicant has to pick the correct answer from several listed with each question. An extensive list of questions, typical of those asked in the examination, is given in the RADIO AMATEUR NEWCOMER. In the code test, you will be required to send and receive messages in plain language, including figures and punctuation marks, at a speed of 13 words per minute (5 characters to the word), for a period of one minute, without mistakes. Both sending and receiving tests are five minutes in length, which gives the applicant an excellent opportunity to send or re-

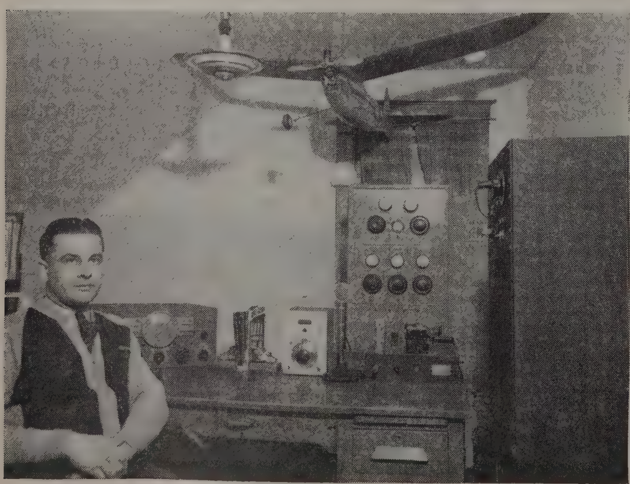


Figure 1.
Amateur station W9RIL, typical of hundreds of medium power "ham" outfits throughout the U.S.A. As usually is the case, the receiving equipment is of commercial manufacture, the transmitting equipment homemade.

ceive the required 65 consecutive characters without error.

If you pass both the code and written tests successfully, you will later receive a combined license from the Commission's offices in Washington. This license, when signed by you, becomes valid. It is a combination operator and station license, one being printed on the reverse side of the other.

Do not confuse the *call areas* (1 to 9) with the *U. S. Radio Districts* (1 to 22). It is rather confusing to the newcomer because amateurs commonly refer to call areas as districts and indicate a station in, for example, the ninth call area as a "ninth district station."

The class B operator license will authorize you to operate c.w. radiotelegraph transmitters (any licensed amateur transmitter, not just your own) in any amateur band or radiophone transmitters in the 160-, 10-, 5-, $2\frac{1}{2}$-, $1\frac{1}{4}$-, and $\frac{3}{4}$-meter bands. You will not be entitled to operate phone in the select 80- and 20-meter bands until you have held your class B license for at least one year and have passed an examination for the class A license.

The Class C License. If you live more than 125 miles air-line distance, from the nearest examining point maintained by the Federal Communications Commission, you may apply for a class C license, the examination for which is given by mail. Other persons allowed to apply for the class C license include (1) applicants who can show a certificate from a reputable physician stating that the applicant is unable because of protracted disability to appear for examination, (2) persons stationed at a camp of the Civilian Conservation Corps, and (3) persons who are in the *regular* military or naval service of the United States at a military post or naval station.

A licensed radiotelegraph operator (other than an amateur operator who himself holds only a class C license) or a regularly employed government radiotelegraph operator must sign the class C applicant's blank in the presence of a notary public, attesting to the applicant's ability to send and receive the continental Morse code at the required speed of 13 words per minute. Do *not* send for class C blanks containing the examinations and questions until you feel you are *ready to take your examination,* as you are not supposed to hold them indefinitely after receiving them.

Holders of class C licenses *may* be required by the Commission to appear at an examining point for a supervised written examination and practical code test at any time during the term of their licenses. This is seldom done except where the Commission has reason to suspect that the applicant would have difficulty in passing the personal examination. For instance, an amateur holding a class C ticket who regularly is heard on the air with a bad note or modulation, or is heard sending always at 8 or 9 words a minute, or repeatedly requests QRS, should not be at all surprised to receive a notice to appear. The class C license will be cancelled if the holder does not appear for examination when called, or if he fails to pass when he does appear.

The privileges granted by the class C license are identical with those of the regular license.

Your operator and station licenses will run concurrently, both expiring together three years from the date of issuance stated on the face of the license. Both may be renewed without examination if an application is filed at least 60 days prior to the indicated date of expira-

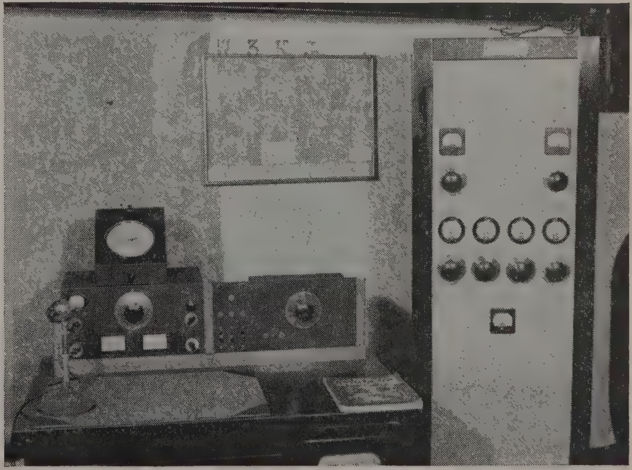


Figure 2.
Another "typical" medium power amateur station, that of W3KJ. Observe the neat simplicity of the installation.

tion and the applicant offers proof that he has communicated via amateur radio with three other amateur stations during the three-month period directly preceding the date of application for renewal.

You may obtain just an operator license (without the station license) if you desire; this will permit you to operate any licensed amateur station. The "station" side of the license will be left blank and you will have no call letters assigned to you. It is not possible to apply for or obtain a station license singly unless you already have an operator license.

Heavy penalties are provided for obtaining an amateur license by fraudulent means, such as by impersonating another person in an examination, copying from notes, books or the like, or misrepresenting the fact of one's U.S. citizenship. Applicants who fail to pass the examination can take it again after two months have passed from the time of the last examination.

There are so many special instances that may arise that no attempt will be made to cover every possible contingency pertaining to the application for and privileges accorded by an amateur license. If you have a special question regarding some point not covered in this book, or which is not clear to you, write to the Inspector-in-Charge of your radio district. Don't guess at the proper interpretation, or take somebody else's word for it; you may get in trouble.

There is one thing you should *not* write to the inspector about and that is the necessity for a license to *transmit*. A transmitter license is absolutely necessary, regardless of power, frequency, or type of emission; there are no exceptions nor special cases.

Before attempting to take the amateur examination, the reader should have a thorough knowledge of the regulations affecting amateur operators and stations. While "memorizing" procedure is not to be recommended when preparing for the *technical* portion of the amateur examination, the best way to prepare for the questions pertaining to regulations is to memorize the pertinent extracts from the communications law, and also the United States amateur regulations, given in Chapter 29. They do not necessarily have to be memorized verbatim, but the applicant must have at his command *all of the information contained therein.*

It is important that the reader clearly understand the distinction between violations of the basic Communications Act of 1934, and violations of the rules and regulations set up under the basic act by the Federal Communications Commission. The former constitutes the more serious offense, and anyone is liable, whether he be an amateur or not. The difficulty some applicants experience with certain questions is in deciding whether a certain offense is a violation of the basic act or a violation of rules set up by our F.C.C. under the act.

Starting Your Study. When you start your study to prepare yourself for the amateur examination, you will probably find that the circuit diagrams, tube characteristic curves, and formulas first appear confusing and difficult of comprehension. However, after putting in a few evenings of study, one becomes sufficiently familiar with basic concepts and fundamentals that the acquisition of further knowledge is not only easy but fascinating.

As it takes considerable time to become proficient at sending and receiving code, it is a good idea to start by interspersing technical

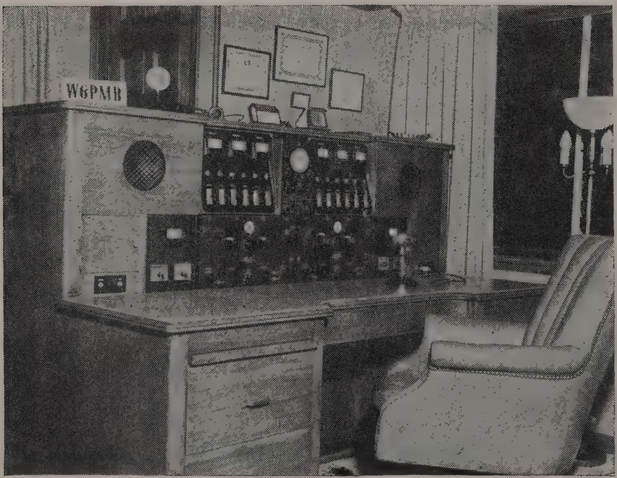


Figure 3.
The receiving position of the prepossessing installation at W6PMB shows that amateur equipment may be made to blend unobtrusively with home furnishings.

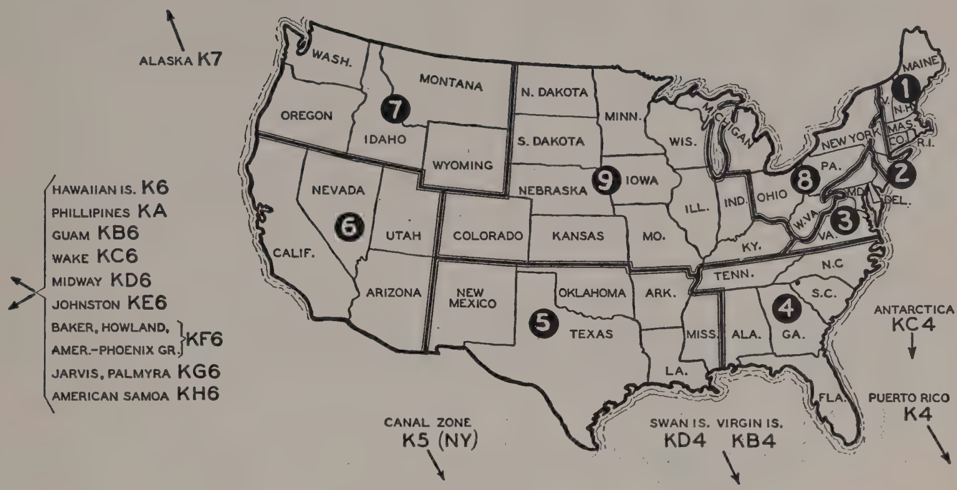


Figure 4.
U.S. call areas, often erroneously referred to as districts. U.S. radio districts (1 to 22) should not be confused with the call areas.

study sessions with short periods of code practice. There are two reasons for this: many short code practice sessions benefit one much more than a fewer number of comparatively longer sessions. Also, it keeps one from getting "stale" while studying theory and regulations by serving as a rest period, thus serving to maintain one's interest. Each kind of study serves as a respite from the other.

You can even start on one of the simpler receivers described in the chapter on *Receiver Construction* if you wish, though at first you will be unable to decipher many of the dots and dashes you pick up on it. However, many interesting hours can be spent listening to the conversations of amateur phone stations. The numerous references to "QSA," "Rig," "Rotary," and other mysterious terms will begin to take on significance.

When you have practiced the code long enough, you will be able to follow the gist of the conversation of slower sending code stations, and fish for "dx." Many stations send slower than "13 per" when working dx. Stations repeat their calls many times when calling "CQ," and one need not have achieved much proficiency to make out their calls and thus determine their location. Granted that it is advisable to start right in with learning the code, you will want to know how to go about mastering it in the shortest possible time with the least amount of effort.

Learning the Code

The applicant for an amateur license must be able to send and receive the Continental code at a speed of 13 words per minute, with an average of 5 characters to the word. Thus 65 characters must be copied consecutively without error in one minute. Similarly, 65 consecutive characters must be transmitted without error in that time. The applicant, however, is given sufficient opportunity to pass this code test, since sending and receiving tests are both five minutes in length. If 65 consecutive characters, at the required rate, are copied correctly, somewhere during the first five-minute period, the applicant may then attempt a transmission. Again, if 65 consecutive characters are sent correctly somewhere during this second period, a passing mark is received.

Failure to pass the code test results in a two-month rest period during which the applicant can improve his mastery of the code; thereafter, he may again appear for another try.

Approximately 30 per cent of the amateur license applicants fail to pass the code examination. It should be expected that nervousness and excitement—at least to some degree—will hinder the applicant's code ability. The best prevention against this is to be able to master the code at a little better than the required speed, under ordinary conditions. Then

THE RADIOTELEGRAPH CODE

A	• —	N	— •
B	— • • •	O	— — —
C	— • — •	P	• — — •
D	— • •	Q	— — • —
E	•	R	• — •
F	• • — •	S	• • •
G	— — •	T	—
H	• • • •	U	• • —
I	• •	V	• • • —
J	• — — —	W	• — —
K	— • —	X	— • • —
L	• — • •	Y	— • — —
M	— —	Z	— — • •

NUMERALS, PUNCTUATION MARKS, ETC.

1	• — — — —	6	— • • • •
2	• • — — —	7	— — • • •
3	• • • — —	8	— — — • •
4	• • • • —	9	— — — — •
5	• • • • •	Ø	— — — — —

INTERNATIONAL DISTRESS SIGNAL	• • • — — — • • •
PERIOD	• — • — • —
COMMA	— — • • — —
INTERROGATION	• • — — • •
QUOTATION MARK	• — • • — •
COLON	— — — • • •
SEMICOLON	— • — • — •
PARENTHESIS	— • — — • —
FRACTION BAR	— • • — •
WAIT SIGN	• — • • •
DOUBLE DASH (BREAK)	— • • • —
ERROR (ERASE) SIGN	• • • • • • • •
END OF MESSAGE	• — • — •
END OF TRANSMISSION	• • • — • —

Figure 5.
Shown above is the Continental code used for all radio communications. The more complicated Morse code is used for land line telegraphic communication within the U.S.A.

a little slowing down due to nervousness will not prove "fatal" during the strain and excitement of the examination. As to the correct method of learning the code, the following is recommended. Unfortunately, no "trick" short cut to learning the code has been found generally successful.

Memorizing the Characters. To memorize the alphabet entails but a few evenings of diligent application. The time required to build up speed will be entirely dependent upon individual ability and regularity of drill, and may take any length of time from a few weeks to many months.

Since code reading requires that individual letters be recognized instantly, any memorizing scheme which depends upon an orderly sequence, such as learning all "dot" letters and all "dash" letters in separate groups, is to be discouraged.

Each letter and figure must be recognized by its *sound* rather than by its appearance. Telegraphy is a system of sound communication, the same as is the spoken word. The letter A, for example, is one short and one long sound in combination, sounding like *did-dah,* and it must be remembered as such, not *dot-dash.*

If you listen to the sound of a letter transmitted slowly by a buzzer and key in the hands of some experienced operator, you will notice how closely the dots resemble the sound *did* and the dashes *dah.*

Before beginning practice with a code-practice set, it is necessary to memorize the whole alphabet perfectly. A good plan is to study only two or three letters a day and to drill with those letters until they become part of your consciousness. Mentally translate each day's letters into their sound equivalent wherever they are seen—on signs, in papers; indoors and outdoors. Tackle two additional letters in the code chart each day, at the same time reviewing the characters already learned.

Avoid memorizing by routine. Be able to sound out any letter immediately without so much as hesitating to think about the letters preceding or following the one in question. Know C, for example, apart from the sequence A, B, C. Skip about among all of the characters learned, and before very long sufficient letters will have been acquired to enable you to spell out simple words to yourself in *"did-dahs."* This is interesting exercise, and for that reason it is good to memorize all of the vowels first, the most common letters next.

Actual code practice should start only when the entire code, including numerals and the few *commonly used* punctuation marks, have been memorized so thoroughly that any letter or figure can be sounded at a moment's notice without hesitation.

Once you have memorized the code thoroughly, you should concentrate on increasing your *receiving* speed. True, if you practice with another newcomer who is learning the code, you will both have to do some sending. But do not attempt to practice *sending* just for the sake of increasing your sending *speed.*

When transmitting on the code practice set to your partner, so that he can get receiving practice, concentrate on the *quality* of your sending, not on your speed. Your partner will appreciate it, and he could not copy you if you "opened up" anyhow. If you want to get a reputation as having an excellent "fist" on the air, just remember that speed alone won't do the trick. Proper execution of your letters and spacing will make much more of an impression. Fortunately, as you get so that you can send evenly and accurately, your sending speed will automatically increase. Remember, try to see how *evenly* you can *send* and how *fast* you can *receive.*

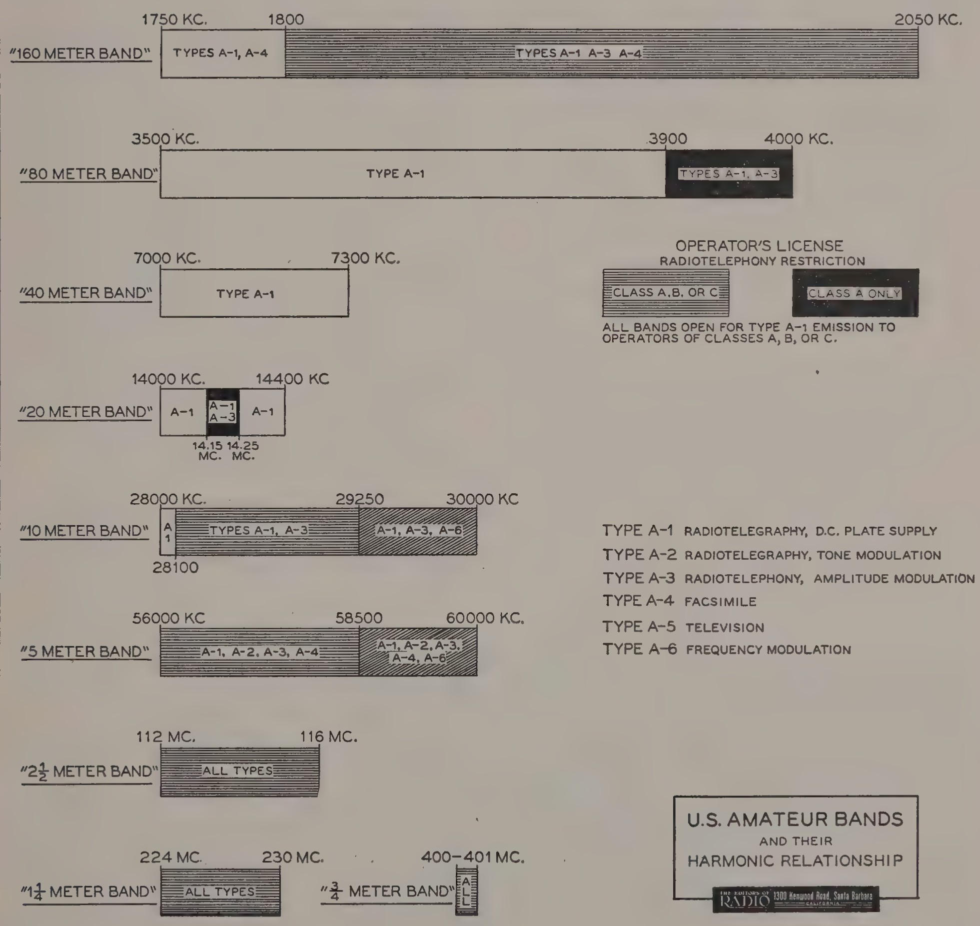


Figure 6.
AMATEUR FREQUENCY BANDS.
This chart is correct as of the time this book goes to press. However, various changes are scheduled for the 1750, 3500, and 7000 kc. bands to provide for government use of certain frequencies shown above as being open to amateurs. Anyone not familiar with the latest changes should request information from his District Inspector before going on the air, as interference to government services is a serious offense.

Because it is comparatively easy to learn to send rapidly, especially when no particular care is given to the quality of the sending, many amateurs who have just received their licenses get on the air and send mediocre code at 20 words a minute when they can barely receive *good* code at 13. While most old timers on the air remember their own period of initiation and are only too glad to be patient and considerate if you tell them you are a beginner, the surest way to incur their scorn is to try to impress them with your "lightning sending" and then request "QRS" when they come back to you at the same speed.

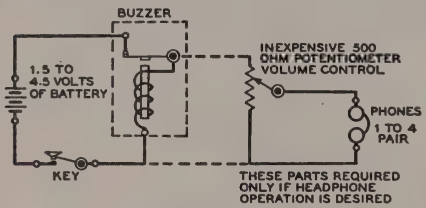


Figure 7.
THE SIMPLEST CODE PRACTICE SET CONSISTS OF A KEY AND BUZZER.
The buzzer is adjusted to give a steady, high pitched "whine." If desired, the phones may be omitted, in which case the buzzer should be mounted firmly on a sounding board. Either crystal or magnetic earphones may be used. Phones should be connected in parallel, not series.

Code Practice Sets. If you don't feel too foolish doing it, you can secure a measure of code practice with the help of a partner by sending "did-dah" messages to each other while riding to work, eating lunch, etc. It is better, however, to use a buzzer or code practice oscillator in conjunction with a regular telegraph key.

As a good key may be considered an investment, it is wise to make a well-made key your first purchase. Regardless of what type code practice set you use, you will need a key, and later on you will need one to key your transmitter. If you get a good key to begin with, you won't have to buy another one later.

The key should be rugged and have fairly heavy contacts. Not only will the key stand up better, but such a key will contribute to the "heavy" type of sending so desirable for radio work. Morse (telegraph) operators use a "light" style of sending and can send somewhat faster when using this light touch. But, in radio work static and interference are often present, and a slightly heavier dot is desirable. If you use a husky key, you will find yourself automatically sending in this manner.

Special types of keys, especially the semi-automatic "bug" type, should be left alone by the beginner. Mastery of the standard type key should come first. The correct manner of using such a key will be discussed later.

To generate a tone simulating a code signal as heard on a receiver, either a mechanical buzzer or an audio oscillator (howler) may be used. The buzzer may be mounted on a sounding board in order to increase the fullness and volume of the tone; or it may be mounted in a cardboard box stuffed with cotton in order to silence it, and the signal fed into a pair of earphones. The latter method makes it possible to practice without annoying other people as much, though the clicking of the key will no doubt still bother someone in the same room.

A buzzer-type code practice circuit is shown in Figure 7. The buzzer should be of good quality or it will change tone during keying; also the contacts on a cheap buzzer will soon wear out. The volume control, however (used only for headphone operation), may be of the least expensive type available, as it will not be subjected to constant adjustment as in a radio receiver. For maximum buzzer and battery life, use the least amount of voltage that will provide stable operation of the buzzer and sufficient volume. Some buzzers operate stably on $1\frac{1}{2}$ volts, while others require more.

A vacuum tube audio oscillator makes the best code practice oscillator, as there is no sound except that generated in the earphones, and the note more closely resembles that of a radio signal. Such a code practice oscillator is diagrammed schematically in Figure 8. The parts are all screwed to a wood board, and connections made to the phones and batteries by means of Fahnestock clips, as illustrated in Figure 9. A single dry cell supplies filament power, and a $4\frac{1}{2}$-volt C battery supplies plate voltage. Both filament and plate current are very low, and long battery life can be expected. The vacuum tube is the biggest item from the standpoint of cost, but it can later be used in a field-strength meter with the same batteries supplying power. Such a device is very handy to have around a station, as it can be used for neutralizing, checking the radiation characteristics of your antenna, etc.

A 1H4, 30, or 1G4G may be used with the same results. The first two are 2-volt tubes, but will work satisfactorily on a 1.5-volt filament battery because of the very small amount of emission required for the low value of plate current drawn. Be sure to get a socket that will accommodate the particular tube you buy.

Oddly, it is important that the audio transformer used *not* be of good quality; if it is, it

may have so much inductance that it will be impossible to get a sufficiently high pitched note. If you buy a new transformer, get the smallest, cheapest one you can buy. The old transformers used in moderately priced sets of 12 years ago are fine for the purpose, and can oftentimes be picked up for a small fraction of a dollar at the "junk parts" stores. The turns ratio is not important; it may be anything between 1.5/1 and 6/1.

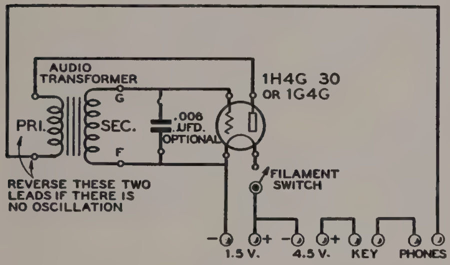


Figure 8.
SIMPLEST TYPE V.T. CODE PRACTICE OSCILLATOR.
Power is furnished by a dry cell and a 4½ volt "C" battery. If the .006-μfd. condenser is omitted, a higher pitched note will result. The note may have too low a pitch even with the condenser omitted, unless the smallest, cheapest audio transformer available is used. The earphones should be of the magnetic type, as plate current must flow through the earphones.

Correct transformer polarity is necessary for oscillation. If oscillation is not obtained, reverse the two wires going to the primary terminals of the transformer.

The tone may be varied by substituting a larger (.025 μfd.) or smaller (.001 μfd.) condenser for the .006 μfd. capacitor shown in the diagram. A lower capacity condenser will raise the pitch of the note somewhat and vice versa. The highest pitch that can be obtained with a given transformer will result when the condenser is left out of the circuit altogether. Lowering the plate voltage to 3 volts will also have a noticeable effect upon the pitch of the note. If the particular transformer you use does not provide a note of a pitch that suits you, the pitch can be altered in this manner.

Using a 1H4G, a standard no. 6 dry cell for filament power, and a 4½-volt C battery for plate power, the oscillator may be constructed for about $2.00, exclusive of key and earphones. The filament battery life will be about 700 hours, the plate battery life considerably more. This set has an advantage over an a.c. operated practice set in that it can be used where there is no 110-volt power available; you can take it on a Sunday picnic if you wish. Also, there is no danger of electrical shock.

The carrier-operated keying monitor described in Chapter 22 also may be used for code practice, and is recommended where loud speaker operation is desired, such as for group practice.

Automatic Code Machines. The two practice sets just described—the buzzer and the v.t. oscillator—are of most value when you have someone with whom to practice. If you are unable to enlist a code partner and have to practice by yourself, the best way to get receiving practice is by use of a set of phonograph code practice records or a tape machine (automatic code-sending machine) with several practice tapes.

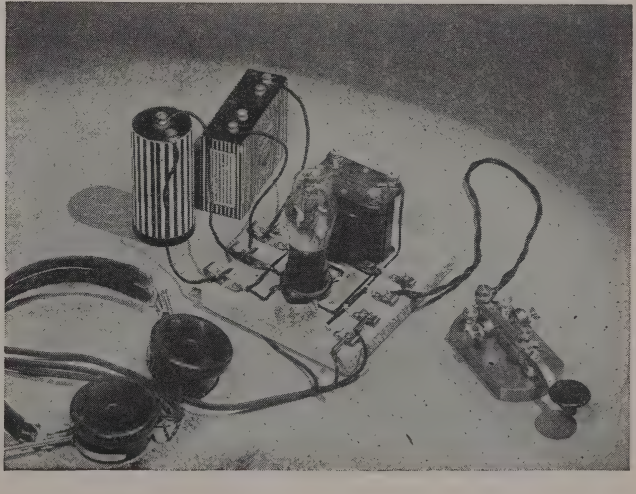


Figure 9.
THE CIRCUIT OF FIGURE 8 IS USED IN THIS BATTERY OPERATED CODE OSCILLATOR.
A tube and audio transformer essentially comprise the oscillator. Fahnestock clips screwed to the baseboard are used to make connections to batteries, key, and phones.

The records are of use only if you have a phonograph whose turntable speed is readily adjustable. The tape machine can be rented by the month for a reasonable fee. Once you can copy close to 10 w.p.m., you can get receiving practice by listening to slow-sending amateurs on your receiver. However, until you can copy around 10 w.p.m., your receiver isn't of much use, and either another operator or a tape machine or code records are necessary for getting receiving practice after you have once memorized the code.

The student must observe the rule always to write down each letter as soon as it is received, never dots and dashes to be translated later. If the alphabet has actually been mastered beforehand, there will be no hesitation from failure to recognize most of the characters unless the transmission speed is too high.

Don't practice too long at one stretch; it does more harm than good. Twenty-five or thirty minutes should be the limit.

Time must not be spent trying futilely to recall a missed letter. Dismiss it and center the attention on the next letter. In order to prevent guessing and to give you equal practice on seldom-used letters such as X, Y, etc., the transmitted material should not be plain language, except perhaps for a few minutes out of each practice period.

During the first practice period, the speed should be such that a substantially solid copy can be made of the entire transmission without strain. Then, in the next period, the speed should be increased slightly to a point where all of the characters can be caught only through conscious effort. When the student becomes proficient at this new speed, another slight increase may be made, progressing in this manner until a speed of about 16 words per minute is attained. The margin of 3 w.p.m. is recommended to overcome the possible excitement factor at examination time. Then, when you take the test you don't have to worry about "jitters" or an "off day."

The speed must not be increased to a new level until the student finally makes solid copy for a 5-minute period at the old level. How frequently increases of speed can be made depends upon individual ability and the number of practice hours. Each increase is apt to prove decidedly disconcerting, but keep in mind the statement by Dr. G. T. Buswell, "You are never learning when you're comfortable."

Using a Key. See Figure 11 for the proper position of the hand, fingers, and wrist when manipulating the telegraph key. The forearm rests naturally on the desk. The knob of the key is grasped lightly with the thumb along the edge and the index and third fingers resting on the top towards the front edge. The hand moves with a free up and down motion, the wrist acting as a fulcrum. The power comes entirely from the arm muscles. The third and index fingers will bend *slightly* during sending, but not because of conscious effort to manipulate the finger muscles. Keep your finger muscles just tight enough to act as a "cushion" for the arm motion and let the slight movement of the fingers take care of itself.

The key spring is adjusted to the individual wrist and should be neither stiff nor "sloppy." Use a moderately stiff tension at first, and gradually lighten it as you get more proficient.

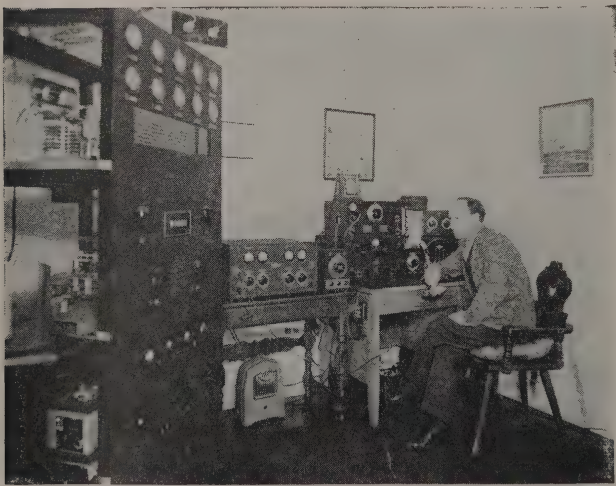


Figure 10.
De luxe high power amateur stations such as that of W6RBQ are not so numerous as are less pretentious installations, but nevertheless there are several hundred throughout the U.S.A.

The separation between the contacts must be the proper amount for the desired speed, being about 1/16 inch for slow speeds and correspondingly closer together (about 1/32 inch) for faster speeds. Avoid extremes in either direction. It is preferable that the key be placed far enough from the edge of the table (about 18 inches) that the elbow can rest on the table.

The characters must be properly spaced and timed, with the dot as the yardstick. A standard dash is three times as long as a dot. The spacing between parts of the same letter is equal to one dot; between letters, three dots; between words, five dots.

This does *not* apply when sending slower than about 10 words per minute for the benefit of someone learning the code and desiring receiving practice. When sending at, say, 5 w.p.m., the individual letters should be made the same as though the sending rate were about 10 w.p.m. except that the spacing between letters and words is greatly exaggerated. The reason for this is obvious. The letter L, for instance, will sound exactly the same at 10 w.p.m. as at 5 w.p.m., and when the speed is increased above 5 w.p.m., the student will not have to become familiar with a new sound (faster combination of dots and dashes). He will merely have to learn the identifying of the *same* sounds without taking so long to do so.

It has been shown that it does not aid a student to identify a letter by sending the individual components of the letter at a speed corresponding to less than 10 words per minute. By sending the letter moderately fast, a longer space can be left between letters for a given code speed, giving the student more time to identify the letter.

When two co-learners have memorized the code and are ready to start sending to each other for practice, it is a good idea to enlist the aid of an experienced operator for the first practice session, so that you will get an idea of what properly formed letters sound like.

When you are practicing with another beginner, don't gloat because you seem to be learning to receive faster than he. It may mean that his *sending* is better than yours. Remember that the quality of sending affects the maximum copying speed of a beginner by as much as 100 per cent. Yes, if the sending is bad enough, a newcomer won't be able to read it at all, even though an old timer may be able to get the general drift of what you are trying to send. A good test for any "fist" is to try it on someone who is just getting to the "13 per" stage.

If You Have Trouble. Should you experience difficulty increasing your code speed after you once have memorized the characters, there is no reason for becoming discouraged. There is no denying that it is more difficult for some people than for others to learn the code, but there is no justification for the contention sometimes made that "some people just can't learn the code." It is not a matter of intelligence; so don't feel ashamed if you seem to experience a little more than the usual difficulty in learning the code. Your reaction time may be a little slower, your coördination not quite so good. If this is the case, remember that you can *still learn the code.* You may never learn to send and receive at 40 w.p.m., but you can learn to send and receive 13 w.p.m. if you have patience and if you refuse to be discouraged by the fact that others seem to pick it up much more rapidly.

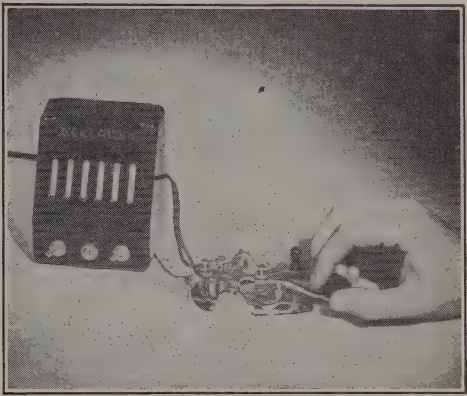


Figure 11.
PROPER POSITION OF FINGERS FOR OPERATING A TELEGRAPH KEY.
The fingers hold the knob and act as a cushion. The hand rests lightly on the key. The muscles of the forearm provide the power, the wrist acting as a fulcrum. The power should not come from the wrist, but rather from the forearm muscles.

While the slow learner can ultimately get his 13 per by following the same learning method if he has *perseverance,* the following system of auxiliary practice will oftentimes prove of great aid in increasing one's speed when progress by the usual method of practice seems to have reached a temporary standstill. All that is required is the usual key-buzzer-headset outfit, plus an extra operator. This last item should be of good quality, guaranteed to

pay proper attention to spacing, and capable of good speed regulation.

Suppose we call the fellow at the key the teacher and the other fellow the student.

Assume the usual positions, but for the moment lay aside the paper and pencil. Instead, the student will read from a duplicate newspaper the *same text the opeartor is sending.*

The teacher is to start sending at a rate just slower than the student's top speed, judged by his last test. This will allow the student to follow accurately each letter as it is transmitted. After a warming-up period of about one minute, the sending speed is to be increased gradually but steadily, and continued for about a period of five minutes. An equal rest period is beneficial before the second session. Speed for the second period should be started at half way between the original starting speed and the speed used at the end of the first period. Follow the same procedure for the second and third practice periods.

At the end of the third *reading* practice period, the student should start copying immediately, using the *same text as before,* at a speed just above the student's previous copying ability. It will be found that one session of the *reading* practice will, for the time being, increase the student's *copying* ability from 10 to 20 per cent. The teacher should watch the student and not increase the sending speed too much above the copying ability of the student, as this brings about a condition of confusion and is more injurious than beneficial.

Code Classes. A number of altruistic amateurs send code practice on schedule once or twice each week, and excellent practice can be obtained after you have bought or constructed your receiver by taking advantage of these sessions. Call letters, time schedules, and frequency in kilocycles of the stations currently sending code practice can be obtained by writing the American Radio Relay League, West Hartford, Conn. Enclose a stamped, self-addressed envelope.

If you live in a large city, the chances are that at least one of the radio clubs or amateur parts stores conducts a free code class. If you inquire around a bit, you usually can discover how to get in on these practice sessions, which also usually provide many interesting social contacts.

CHAPTER TWO

Fundamental Radio and Electrical Theory

Constitution of Matter. All matter is made up of approximately 92 fundamental constituents commonly called *elements*. These elements can exist either in the free state such as iron, oxygen, carbon, copper, tungsten, and aluminum, or in chemical unions commonly called *compounds*. The smallest unit which still retains all the original characteristics of an element is the *atom*.

Combinations of atoms, or subdivisions of compounds, result in another fundamental unit, the *molecule*. The molecule is the smallest unit of any compound. All reactive elements when in the gaseous state also exist in the molecular form, made up of two or more atoms. The nonreactive or noble gaseous elements helium, neon, argon, krypton, xenon, and radon are the only gaseous elements that ever exist in a truly atomic state.

The Atom. An atom is an extremely small unit of matter—there are literally billions of them making up so small a piece of material as a speck of dust. But to understand the basic theory of electricity and hence of radio, we must go further and divide the atom into its main components, a positively charged nucleus and a cloud of negatively charged particles that surround the nucleus. These particles, swirling around the nucleus in elliptical orbits at an incredible rate of speed, are called orbital electrons. It is upon the behavior of these electrons that depends the study of electricity and radio, as well as allied sciences. Actually it is possible to subdivide the nucleus of the atom into other particles: the proton, nuclear electron, negatron, positron, and neutron; but this further subdivision can be left to quantum mechanics and atomic physics. As far as radio theory is concerned it is only necessary for the reader to think of the normal atom as being composed of a nucleus having a net positive charge that is exactly neutralized by the one or more orbital electrons surrounding it.

The atoms of different elements differ in respect to the charge on the positive nucleus and in the number of electrons revolving around this charge. They range all the way from hydrogen, having a net charge of one on the nucleus and one orbital electron, to uranium with a net charge of 92, and 92 orbital electrons. The number of orbital electrons is called the *atomic number* of the element.

From the above it must not be thought that the electrons revolve in a haphazard manner around the nucleus. Rather, the electrons in an element having a large atomic number are grouped into "shells" having a definite number of electrons. In the first shell there is room for only 2 electrons; in the next, 2; the next, 6; then 2, 6, 10, 2, 6, 10, etc., until a total of 92 electrons can be accommodated in the heaviest atom, that of uranium. The only atoms in which these shells are completely filled are those of the inert or noble gases mentioned before; all other elements have one or more uncompleted shells of electrons. If the uncompleted shell is nearly empty, the element is *metallic* in character, being most metallic when there is only one electron in the outer shell. If the incomplete shell lacks only one or two electrons, the element is usually *non-metallic*. Elements with a shell about half completed will exhibit both non-metallic and metallic character; carbon, silicon, and arsenic are examples of this type of element.

In metallic elements these outer-shell electrons are rather loosely held. Consequently, there is a continuous helter-skelter movement of these electrons and a continual shifting from one atom to another. The electrons which move about in a substance are called *free electrons*, and it is the ability of these electrons to drift from atom to atom which makes possible the *electric current*.

Electromotive Force: Potential Difference. The free electrons in a conductor move constantly about and change their position in a haphazard manner. To produce a drift of electrons along a wire, or electric current, it is necessary that there be a difference in pressure or potential between the two ends of the wire. This *potential difference* can be produced by connecting a battery to the ends of the wire.

As will be explained later, there is an excess of electrons at the negative terminal of a battery and a deficiency of electrons at the positive terminal, due to chemical action. When the battery is connected to the wire the deficient atoms at the positive terminal attract free electrons from the wire in order to become neutral. The attracting of electrons continues through the wire, and finally the excess electrons at the negative terminal of the battery are attracted by the positively charged atoms at the end of the wire. The same result would be obtained if the wire were connected between the terminals of a generator.

Thus it is seen that a potential difference is the result of a difference in the number of electrons between the two (or more) points in question. The force or pressure due to a potential difference is termed the *electromotive force*, usually abbreviated *e.m.f.* or *E.M.F.* It is expressed in units called *volts*.

The Electric Current. The drift of electrons along a conductor due to the application of an electromotive force constitutes an electric current. This drift is in addition to the irregular movement of the electrons. However, it must not be thought that each free electron travels from one end of the circuit to the other. On the contrary, each free electron travels only a short distance before colliding with an atom; this collision generally knocking off one or more electrons from the atom, which in turn move a short distance and collide with other atoms, knocking off other electrons. Thus, in the general drift of electrons along a wire carrying an electric current, each electron travels only a short distance and the excess of electrons at one end and the deficiency at the other are balanced by the source of the e.m.f. When this source is removed the state of normalcy returns; there is still the rapid interchange of free electrons between atoms, but there is no general trend in either one direction or the other.

There are two units of measurement associated with current, and they are often confused. The *rate of flow* of electricity is stated in *amperes*. The unit of *quantity* is the *coulomb*. A coulomb is equal to 6.28×10^{18} electrons, and when this quantity of electrons flows by a given point in every second, a current of one ampere is said to be flowing. An ampere is equal to one coulomb per second; a coulomb is, conversely, equal to one ampere-second. Thus we see that *coulomb* indicates *amount*, and *ampere* indicates rate of *flow*.

Many textbooks speak of current flow in the *conventional* sense; that is, from the positive terminal of the e.m.f. source through the conductor to the negative terminal. Nevertheless, it has long been an established fact that the current flow in a metallic conductor is the *electronic* flow from the negative terminal of the source of voltage through the conductor to the positive terminal. This is easily seen from a study of the foregoing explanation of the subject. The only exceptions to the electronic direction of flow occur in gaseous and electrolytic conductors where the flow of positive ions toward the cathode or negative electrode constitutes a positive flow in the opposite direction to the electronic flow.

Resistance. The flow of current in a material depends upon the ease with which electrons can be detached from the atoms of the material and upon its molecular structure. In other words, the easier it is to detach electrons from the atoms the more free electrons there will be to contribute to the flow of current, and the fewer collisions that occur between free electrons and atoms the greater will be the total electron flow.

This opposition to a steady electron flow is called the *resistance* of a material, and is one of its physical properties. The resistance of a uniform length of a given substance is directly proportional to its length and specific resistance, and inversely proportional to its cross-sectional area. A wire with a certain resistance for a given length will have twice as much resistance if the length of the wire is doubled. For a given length, doubling the cross-sectional area of the wire will *halve* the resistance.

The resistance also depends upon temperature, increasing with increases in temperature for most substances (including most metals), due to increased electron acceleration and hence a greater number of impacts between electrons and atoms. However, in the case of some substances such as carbon and glass the temperature coefficient is negative, which means that the resistance decreases as the temperature increases. This is also true of electrolytes. The temperature may be raised by the external application of heat, or by the flow of the current itself. In the latter case, this is due to the fact that heat is generated when the electrons and atoms collide. (See *Heating Effect.*)

The unit of resistance is the *ohm*. Every substance has a *specific resistance*, usually expressed as *ohms per mil-foot*, which is deter-

mined by the molecular structure and temperature of the material. A mil-foot is a piece of material one circular mil in area and one foot long.

Conductors and Insulators. In the molecular structure of many materials such as glass, porcelain, and mica all electrons are tightly held within their orbits and there are comparatively few free electrons. This type of substance will conduct an electric current only with great difficulty and is known as an *insulator*. An insulator is said to have a high electrical *resistance*.

On the other hand, materials that have a large number of free electrons are known as *conductors*. Most metals, those elements which have only one or two electrons in their outer shell, are good conductors. Silver, copper, and aluminum, in that order, are the best of the common conductors and are said to have the greatest *conductivity*, or lowest resistance to the flow of an electric current.

Fundamental Electrical Units. These units are the *volt*, the *ampere*, and the *ohm*. They were mentioned in the preceding paragraph, but were not completely defined.

The fundamental unit of *current*, or *rate of flow* of electricity is the ampere. A current of one ampere will deposit silver from a specified solution of silver nitrate at a rate of 1.118 milligrams per second.

The international standard for the ohm is the resistance offered by a column of mercury at 0° C., 14.4521 grams in mass, of constant cross-sectional area and 106.300 centimeters in length. The expression *megohm* (1,000,000 ohms) is also sometimes used when speaking of very large values of resistance.

A volt is the voltage that will produce a current of one ampere through a resistance of one ohm. The standard of electromotive force is the Weston cell which at 20° C. has a potential of 1.0183 volts across its terminals. This cell is used only for reference purposes, since only an infinitesimal amount of current may be drawn from it without disturbing its characteristics.

Ohm's Law. The relationship between the electromotive force (voltage), the flow of current (amperes), and the resistance which impedes the flow of current (ohms), is very clearly expressed in a simple but highly valuable law known as *Ohm's law*. This law states that *the current in amperes is equal to the voltage in volts divided by the resistance in ohms*. Expressed as an equation:

$$I = \frac{E}{R}$$

If the voltage (E) and resistance (R) are known, the current (I) can be readily found. If the voltage and current are known, and the resistance is unknown, the resistance (R) is equal to $\frac{E}{I}$. When the voltage is the unknown quantity, it can be found by multiplying $I \times R$. These three equations are all secured from the original by simple transposition. The expressions are here repeated for quick reference:

$$I = \frac{E}{R} \qquad R = \frac{E}{I} \qquad E = IR$$

where I is the current in amperes,
R is the resistance in ohms,
E is the electromotive force in volts.

Applications of Ohm's Law. All electrical circuits fall into one of three classes: series circuits, parallel circuits, and series-parallel circuits. A series circuit is one in which the current flows in a single continuous path and is of the same value at every point in the circuit. In a parallel circuit there are two or more current paths between two points in the circuit, as shown in Figure 2. Here the current divides at A, part going through R_1 and part through R_2, and combines at B to return to the battery. Figure 4 shows a series-parallel circuit. There are two paths between points A and B as in the parallel circuit, and in addition there are two resistances in series in each branch of the parallel combination. Two other examples of series-parallel arrangements appear in Figure 5. The way in which the current splits to flow through the parallel branches is shown by the arrows.

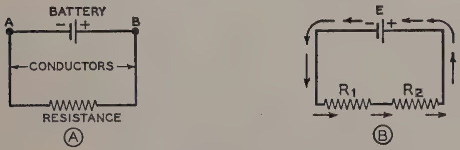


Figure 1.
SIMPLE SERIES CIRCUITS.

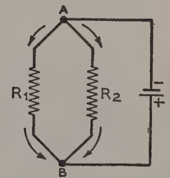


Figure 2.
SIMPLE PARALLEL CIRCUITS.

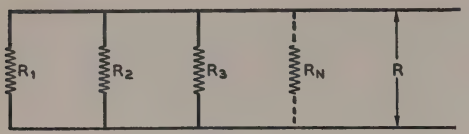
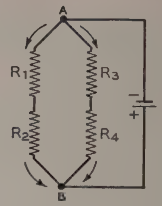


Figure 3.
RESISTORS IN PARALLEL.

Figure 4.
SERIES-PARALLEL
CIRCUIT.



In every circuit, each of the parts has some resistance; the batteries or generator, the connecting conductors, and the apparatus itself. Thus, if each part has some resistance, no matter how little, and a current is flowing through it, there will be a voltage drop across it. In other words, there will be a potential difference between the two ends of the circuit element in question. This drop in voltage is equal to the product of the current and the resistance, hence it is called the *IR* drop.

The source of voltage has an *internal* resistance, and when connected into a circuit so that current flows, there will be an IR drop in the source just as in every other part of the circuit. Thus, if the terminal voltage of the source could be measured in a way that would cause no current to flow, it would be found to be less than the voltage measured when a current flows by the amount of the IR drop in the source. The voltage measured with no current flowing is termed the *no load* voltage; that measured with current flowing is the *load* voltage. It is apparent that a voltage source having a low internal resistance is most desirable, in order that the internal IR drop will be as small as possible, thereby making the terminal voltage more nearly equal to the no load voltage.

Resistances in Series. The current flowing in a series circuit is equal to the voltage impressed divided by the *total* resistance in the circuit. Suppose that the resistance in Figure 1A has a value of 10 ohms and the internal resistance of the battery is 0.5 ohm. The voltage applied between points A and B is 10 volts. Find the current flowing in the circuit. Since the same current flows through every part of the circuit, it is merely necessary to add all the individual resistances to obtain the total resistance. Expressed as a formula:

$$R_{total} = R_1 + R_2 + R_3 + \ldots + R_N$$

Of course, if the resistances happened to be all of the same value, the total resistance would be the resistance of one multiplied by the number in the circuit. In the problem, the total resistance would be 10 ohms plus 0.5 ohm, or 10.5 ohms. From Ohm's law:

$$I = \frac{E}{R}$$

$$E = 10 \text{ volts,}$$
$$R = 10.5 \text{ ohms,}$$

$$\text{and } I = \frac{10}{10.5} \text{ or } 0.95 \text{ ampere.}$$

Knowing the current and the internal resistance of the battery, it is a simple matter to find the IR drop in the battery and the no-load voltage.

$$E = I \times R \text{ or } IR$$
$$I = 0.95 \text{ ampere}$$
$$R = 0.50 \text{ ohm}$$
$$E = 0.95 \times 0.50 = 0.475 \text{ volt}$$

The IR drop in the battery is therefore 0.475 volt. The no-load voltage of the battery would be 10 volts plus 0.475 volt, or 10.475 volts.

As another example, suppose that the voltage at the plate of a vacuum tube is 50 volts, the plate current is 5 milliamperes, and the plate resistor is 50,000 ohms. What voltage would the power supply have to deliver under these conditions? A milliampere (abbreviated *ma.*) is a thousandth of an ampere, so to convert milliamperes to amperes the decimal point is moved three places to the left. Hence 5 ma. = 0.005 ampere. The voltage drop in the plate resistor is found as follows:

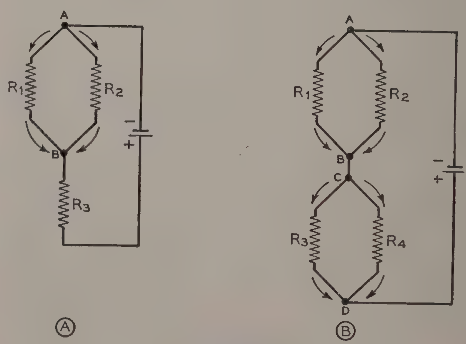


Figure 5.
TWO OTHER SERIES-PARALLEL
CIRCUITS.

$E = IR$
$I = 0.005$ amp.
$R = 50{,}000$ ohms
$E = 0.005 \times 50{,}000 = 250$ volts drop across the resistor

Therefore, if 50 volts are needed at the plate of the tube, the supply voltage must be 50 plus 250, or 300 volts.

Resistances in Parallel. Consider two resistors, one of 100 ohms and one of 10 ohms, connected in parallel as in Figure 2, with a voltage of 10 volts applied across the combination. The same voltage is across each resistor, so the current through each can be easily calculated.

$$I = \frac{E}{R}$$

$E = 10$ volts
$R = 100$ ohms

$$I = \frac{10}{100} = 0.1 \text{ ampere}$$

$E = 10$ volts
$R = 10$ ohms

$$I = \frac{10}{10} = 1.0 \text{ ampere}$$

Until it divides at A, the entire current of 1.1 amperes is flowing through the conductor from the battery, and again from B through the conductor to the battery. Since this is more current than flows through the smaller resistor it is evident that the resistance of the parallel combination must be less than 10 ohms, the resistance of the smaller resistor. We can find this value by applying Ohm's law.

$$R = \frac{E}{I}$$

$E = 10$ volts
$I = 1.1$ amperes

$$R = \frac{10}{1.1} = 9.09 \text{ ohms}$$

The resistance of the parallel combination is 9.09 ohms.

Mathematically, we can derive a simple formula for finding the effective resistance of two resistors connected in parallel. This formula is:

$$R = \frac{R_1 \times R_2}{R_1 + R_2},$$

where R *is the unknown resistance,*
R_1 *is the resistance of the first resistor,*
R_2 *is the resistance of the second resistor.*

If the effective value required is known, and it is desired to connect one unknown resistor in parallel with one of known value, to obtain this unknown value the following transposition of the above formula will simplify the problem:

$$R_2 = \frac{R_1 \times R}{R - R_1},$$

where R *is the effective value required,*
R_1 *is the known resistor,*
R_2 *is the value of the unknown resistance necessary to give* R *when in parallel with* R_1.

The resultant value of placing a number of unlike resistors in parallel is equal to the reciprocal of the sum of the reciprocals of the various resistors. This can be expressed as:

$$R = \frac{1}{\frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3} + \dots \frac{1}{R_n}}$$

The effective value of placing any number of unlike resistors in parallel can be determined from the above formula. However, it is commonly used only when there are three or more resistors under consideration, since the simplified formula given in the following paragraph is more convenient when only two resistors are being used.

When two or more resistances of the same value are placed in parallel, the effective resistance of the paralleled resistors is equal to the value of one of the resistors divided by the number of resistors in parallel.

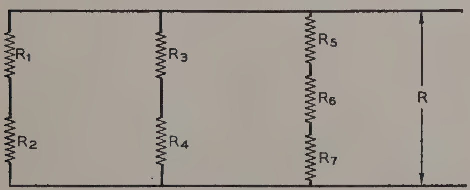


Figure 6.
RESISTORS IN SERIES PARALLEL.

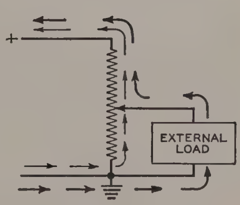


Figure 7.
Indicating flow of current through a tapped voltage divider to an external load.

The effective value of resistance of two or more resistors connected in parallel is *always* less than the value of the lowest resistance in the combination. It is well to bear this simple rule in mind, as it will assist greatly in approximating the value of paralleled resistors.

Shunts. When a voltage is applied to a circuit consisting of two or more resistances in parallel the resulting current divides itself among the paths in inverse proportion to the resistance of each path. With respect to one of the elements those connected in parallel with it are said to *shunt* it.

An example of a shunt which is of particular interest is the use of a resistance to shunt an ammeter or milliammeter (a device for measuring current) so that part of the current in the circuit will be bypassed around the meter. By this means the range of a meter may be greatly extended. Multiplying the range by powers of 10 makes it possible to use the original calibration scale without having to perform mental calculations in taking readings.

To calculate the amount of resistance required in a given case, the basic form of Ohm's law can be used. However, the following formula (derived from Ohm's law) simplifies the calculations:

$$R = \frac{R_m \times I_m}{I - I_m},$$

where R = resistance of shunt in ohms,
R_m = resistance of meter in ohms,
I_m = full scale current for meter,
I = full scale current for new calibration.

Resistances in Series-Parallel. To find the total resistance of several resistors connected in series-parallel, it is usually easiest to apply either the formula for series resistors or the parallel resistor formula first, in order to reduce the original arrangement to a simpler one. For instance, in Figure 4 the series resistors should be added in each branch, then there will be but two resistors in parallel to be calculated. Similarly in Figure 6, although here there will be three paralleled resistors after adding the series resistors in each branch. In Figure 5 the paralleled resistors should be reduced to the equivalent series value, and then the series resistance values can be added.

Resistances in series-parallel can be solved by combining the series and parallel formulas into one similar to the following (refer to Figure 6):

$$R = \frac{1}{\dfrac{1}{R_1 + R_2} + \dfrac{1}{R_3 + R_4} + \dfrac{1}{R_5 + R_6 + R_7}}$$

However, this method is usually more complicated than that mentioned above.

Bleeder Resistors. Resistors are often connected across the output terminals of power supplies in order to bleed off a constant value of current or to serve as a constant fixed load. The regulation of the power supply is thereby improved, and the voltage is maintained at a more or less constant value, regardless of load conditions. When the load is entirely removed from a power supply, the voltage may rise to such a high value as to ruin the filter condensers.

The amount of current which can be drawn from a power supply depends upon the current rating of the particular power transformer in use. If a transformer will carry a maximum safe current of 100 milliamperes, and if 75 milliamperes of this current is required for operation of a radio receiver, there remains 25 milliamperes of current available which can be wasted in the bleeder resistor.

An example for calculating bleeder resistor values is as follows: The power supply delivers 300 volts. The power transformer can safely supply 75 milliamperes of current, of which 60 milliamperes will be required for the receiver. The problem is to find the correct value of resistance to give a bleeder current of 15 milliamperes. Ohm's law gives the solution:

$$R = \frac{E}{I} = \frac{300}{.015} = 20{,}000 \text{ ohms.}$$

(15 milliamperes is equivalent to .015 ampere.) Therefore, it is seen that the bleeder resistor should have a resistance of 20,000 ohms.

Voltage Dividers. A voltage divider is exactly what its name implies: a resistor or a series of resistors connected across a source of voltage from which various lesser values of voltage may be obtained by connection to various points along the resistor.

A voltage divider serves a most useful purpose in a radio receiver, transmitter or amplifier, because it offers a simple means of obtaining plate, screen, and bias voltages of different values from a common power supply source. It may also be used to obtain very low voltages of the order of .01 to .001 volt with a high degree of accuracy, even though a means of measuring such voltages is lacking. The procedure for making these measurements can best be given in the following example:

Assume that an accurately calibrated voltmeter reading from 0 to 150 volts is available, and that the source of voltage is exactly 100 volts. This 100 volts is then impressed through a resistance of exactly 1,000 ohms. It will, then, be found that the voltage along

various points on the resistor, with respect to the grounded end, will be exactly proportional to the resistance at that point. From Ohm's law, the current would be 0.1 ampere; this current remains unchanged since the original value of resistance (1,000 ohms) and the voltage source (100 volts) are unchanged. Thus, at a 500-ohm point on the resistor (half its entire resistance), the voltage will likewise be halved or reduced to 50 volts.

The equation ($E = I \times R$) gives the proof: $E = 500 \times 0.1 = 50$. At the point of 250 ohms on the resistor, the voltage will be one-fourth the total value, or 25 volts ($E = 250 \times 0.1 = 25$). Continuing with this process, a point can be found where the resistance measures exactly 1 ohm and where the voltage equals 0.1 volt. It is, therefore, obvious that if the original source of voltage and resistance can be measured, it is a simple matter to predetermine the voltage at any point along the resistor, provided that the current remains constant, and provided that no current is taken from the tap-on point unless this current is taken into consideration.

Design of Voltage Dividers. Proper design of a voltage divider for any type of radio equipment is a relatively simple matter. The first consideration is the amount of bleeder current to be drawn, which is dictated largely by the examples previously given. In addition, it is also necessary that the desired voltage and the exact current at each tap on the voltage divider be known.

Figure 7 illustrates the flow of current in a simple voltage divider and load circuit. The light arrows indicate the flow of bleeder current, while the heavy arrows indicate the flow of the load current. The design of a combined bleeder resistor and voltage divider, such as is commonly used in radio equipment, is illustrated in the following example.

A power supply delivers 300 volts and is conservatively rated to supply all needed current for the receiver and still allow a bleeder current of 10 milliamperes. The following voltages are wanted: 75 volts at 2 milliamperes for the detector tube, 100 volts at 5 milliamperes for the screens of the tubes, and 250 volts at 20 milliamperes for the plates of the tubes. The required voltage drop across R_1 is 75 volts, across R_2 25 volts, across R_3 150 volts, and across R_4 it is 50 volts. These values are shown in the diagram of Figure 8. The respective current values are also indicated. Applying Ohm's law:

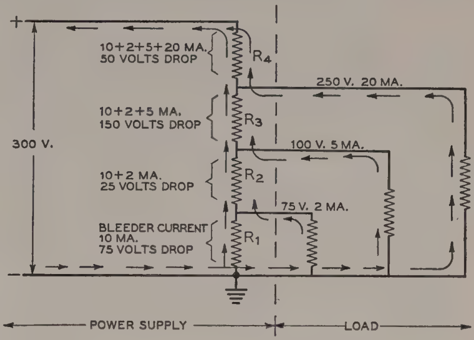


Figure 8.
Combined bleeder resistor and voltage divider.

$$R_1 = \frac{E}{I} = \frac{75}{.01} = 7{,}500 \text{ ohms.}$$

$$R_2 = \frac{E}{I} = \frac{25}{.012} = 2{,}083 \text{ ohms.}$$

$$R_3 = \frac{E}{I} = \frac{150}{.017} = 8{,}823 \text{ ohms.}$$

$$R_4 = \frac{E}{I} = \frac{50}{.037} = 1{,}351 \text{ ohms.}$$

$$R_{Total} = 7{,}500 + 2{,}083 + 8{,}823 + 1{,}351 = 19{,}757 \text{ ohms.}$$

A 20,000-ohm resistor with three sliding taps will be of the approximately correct size, and would ordinarily be used because of the difficulty in securing four separate resistors of the exact odd values indicated, and because no adjustment would be possible to compensate for any slight error in estimating the probable currents through the various taps.

When the sliders on the resistor once are set to the proper point, as in the above example, the voltages will remain constant at the values shown as long as the current remains a constant value.

Disadvantages of Voltage Dividers. One of the serious disadvantages of the voltage divider becomes evident when the current drawn from one of the taps changes. It is obvious that the voltage drops are interdependent and, in turn, the individual drops

are in proportion to the current which flows through the respective sections of the divider resistor. The only remedy lies in providing a heavy steady bleeder current in order to make the individual currents so small a part of the total current that any change in current will result in only a slight change in voltage. This can seldom be realized in practice because of the excessive values of bleeder current which would be required.

When a power supply is used for C-bias service, still another factor must be taken into consideration. Rectified grid current from the class B or C stages will flow through the divider in the same direction as the bleeder current. If this grid current changes, the voltage applied to the grid will also correspondingly change. Adjustments of a C-bias supply should be made while the amplifier draws its proper amount of grid current; otherwise, the C-bias resistor setting will be greatly in error. Heavy bleeder currents are thus required for C-bias supplies, especially where the grid current is changing and the bias must remain constant, as in certain types of 'phone transmitters.

Since the grid current in a C-bias supply flows from the tap on the divider to ground, and in the same direction as the bleeder current, it is important to remember that in this case the regulation is in the opposite direction from the case where power is being taken from a tap on the divider. In other words, the greater the grid current that is flowing through the bleeder, the *higher* will be the voltage at this tap on the divider—and, for that matter, at all other taps in the same divider.

Filaments or Heaters in Series and Parallel. Vacuum tube filaments or heaters are rated by the voltage that should be applied and the current that should flow through them. Tubes having the same filament or heater voltage *and current* ratings may be connected in series across a voltage supply having an output equal to the voltage required for one tube times the number of tubes in series. If the source voltage is less or more than the product of the number of tubes and the voltage required for one tube, the individual tubes will receive less or more than their rated voltage.

If the current requirements of the tubes in series are not all the same, it will be necessary to connect resistors across those tubes which take less current, in order that the total current through the parallel circuit will equal the current rating of the other tubes. This will be made clear by the following example:

A 6F6 and a 6L6 are to be operated from a heater voltage supply which delivers 12.6 volts. The *tube tables* show that a type 6F6 tube draws 0.7 ampere at 6.3 volts, while the type 6L6 tube draws 0.9 ampere at the same voltage. Since each tube requires 6.3 volts, which is half the voltage of the source, it is clear that the two tubes should be connected in series across the source. However, in order to make the voltage drop across each tube equal 6.3 volts, it will be necessary to connect a resistor in parallel with the heater of the 6F6 tube. This resistor must pass 0.2 ampere, the difference between the heater currents of the two tubes. The proper resistor value is found by Ohm's law:

$$R = \frac{E}{I} = \frac{6.3}{0.2} = 31.5 \text{ ohms.}$$

By connecting a 31.5-ohm resistor in shunt with the 6F6 tube heater, the effective resistance of the combination is equal to the resistance of the 6L6 tube heater, and the voltage drop across the combination will be the same as across the 6L6 tube.

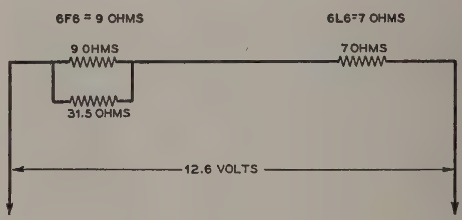


Figure 9.
Obtaining the proper filament voltage drop across each of a pair of dissimilar tubes by means of a resistor across the heater of the one drawing the least amount of current.

If these tubes are connected in series without precautionary measures, the total resistance of the two will be 16 ohms (9 + 7). A potential of 12.6 volts will pass a current of 0.787 ampere through this value of 16 ohms. The drop across each separate resistor is found from Ohm's law, as follows: $9 \times 0.787 = 7.083$ volts, and $7 \times 0.787 = 5.4$ volts. Thus, it is seen that neither tube will have the correct voltage drop.

The resistance of a tube heater or filament should never be measured when cold, because the resistance will be only a fraction of the resistance present when the tube functions at proper heater or filament temperature. The resistance can be calculated satisfactorily by using the current and voltage ratings given by the manufacturer.

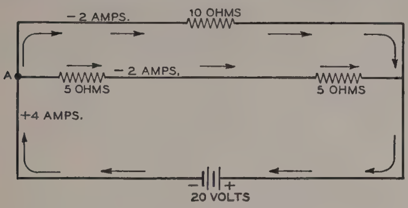


Figure 10-A.
ILLUSTRATING KIRCHHOFF'S FIRST LAW.
The current flowing toward "A" is equal to the current flowing away from "A".

Kirchhoff's Laws. Ohm's law is all that is necessary to calculate the values in simple circuits, such as the preceding examples; but in more complex problems, involving more than one voltage in the same closed circuit, the use of *Kirchhoff's laws* will greatly simplify the calculations. These laws are actually merely rules for applying Ohm's law.

The first law states that at any point in a circuit the current flowing toward the point is equal to the current flowing from it. In other words, if currents flowing to the point are considered positive and those flowing from the point are considered negative, their sum—taking signs into account—is zero. Such a sum is known as an *algebraic sum.*

Figure 10A illustrates this first law. It is readily seen that 4 amperes flow toward point A, and 2 amperes flow away through the two 5-ohm resistors in series, while the remaining 2 amperes flow away through the 10-ohm resistor. Thus, there are 4 amperes flowing to point A and 4 amperes flowing away from the point. If R is the effective resistance of the network, R_1 = 10 ohms, R_2 = 5 ohms, R_3 = 5 ohms, and E = 20 volts, we can set up the following equation:

$$\frac{E}{R} - \frac{E}{R_1} - \frac{E}{R_2 + R_3} = 0$$

$$\frac{20}{5} - \frac{20}{10} - \frac{20}{5 + 5} = 0$$

$$4 - 2 - 2 = 0$$

Kirchhoff's second law states that in any closed path in a network the sum of the IR drops must equal the sum of the applied e.m.f.s, or, the algebraic sum of the IR drops and the applied e.m.f.s in any closed path in a network is zero. The applied e.m.f.s are considered positive, while IR drops taken in the direction of current flow (including the internal drop of the source) are considered negative.

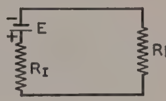


Figure 10-B.
To dissipate the greatest amount of power in the load, R_L should be equal to R_I.

A simple example illustrating the second law is given in Figure 10B. If R_1 = 25 ohms, R_2 = 30 ohms, and E = 15 volts, then the current:

$$I = \frac{E}{R_1 + R_2},$$

$$\text{or } I = \frac{15}{25 + 30} = 0.27 \text{ ampere.}$$

The resistance drops are:

$$IR_1 = 0.27 \times 25 = 6.82 \text{ volts,}$$
$$\text{and } IR_2 = 0.27 \times 30 = 8.18 \text{ volts.}$$

Thus $+ 15 - 6.82 - 8.18 = 0$.

Power in Resistive Circuits. In order to cause electrons to flow through a conductor, constituting a current flow, it is necessary to apply an electromotive force (voltage) across the circuit. Less power is expended in creating a small current flow than in creating a large one, so it is necessary to have a unit of power as a reference.

The unit of electrical power is the *watt,* which is the amount of power used when an e.m.f. of 1 volt forces a current of 1 ampere through a circuit. The power in a resistive circuit is equal to the product of the voltage applied across, and the current flowing in, a given circuit. Hence: P (watts) = E (volts) × I (amperes).

Since it is often convenient to express power in terms of the resistance of the circuit and the current flowing through it, a substitution of IR for E (E = IR) in the above formula gives: $P = IR \times I$ or $P = I^2R$. In terms of voltage and resistance, $P = E^2/R$. Here, $I = E/R$ and when this is substituted for I the original formula becomes $P = E \times E/R$ or $P = E^2/R$. To repeat these three expressions:

$$P = EI,\ P = I^2R, \text{ and } P = E^2/R,$$

where P is the power in watts,
E is the electromotive force in volts, and
I is the current in amperes.

To apply the above equations to a typical problem: The voltage drop across a cathode resistor in a power amplifier stage is 50 volts; the plate current flowing through the resistor is 150 milliamperes. The number of watts the resistor will be required to dissipate is found from the formula: P = EI, or 50 × .150 =

7.5 watts (.150 amperes is equal to 150 milliamperes). From the foregoing it is seen that a 7.5-watt resistor will safely carry the required current, yet a 10- or 20-watt resistor would ordinarily be used to provide a safety factor.

In another problem, the conditions being similar to those above, but with the resistance and current being the *known* factors, the solution is obtained as follows: $P = I^2R = .0225 \times 333.33 = 7.5$. If only the voltage and resistance are known, $P = E^2/R = 2500/333.33 = 7.5$ watts. It is seen that all three equations give the same result; the selection of the particular equation depends only upon the known factors.

Heating Effect. Heat is generated when a source of voltage causes a current to flow through a resistor (or, for that matter, through any conductor). As explained earlier, this is due to the fact that heat is given off when free electrons collide with the atoms of the material. More heat is generated in high resistance materials than in those of low resistance, since the free electrons must strike the atoms harder to knock off other electrons. The heating of the conductor increases the velocity of the free electrons, causing more and harder collisions. As the heating effect is a function of the current flowing and the resistance of the circuit, the power expended in heat is given by the second formula: $P = I^2R$.

Load Matching. To develop the maximum power in the load upon a source of e.m.f., it is necessary to make the resistance (or impedance) of the load equal to the internal resistance (or impedance) of the source. This can best be illustrated by Figure 10B, assuming R_I is the internal resistance of the source and has a value of 1 ohm, while the source E has a no-load voltage of 2 volts. If the load resistance R_L is also 1 ohm, the current is:

$$I = \frac{E}{R_I + R_L} = \frac{2}{1 + 1} = 1 \text{ ampere.}$$

The total power dissipated is:

$$P = EI = 2 \times 1 = 2 \text{ watts,}$$

which is divided equally between the source and the load.

If R_L is 2 ohms the current is:

$$I = \frac{2}{1 + 2} = 0.67 \text{ ampere,}$$

and the total power dissipated is:

$$P = 2 \times 0.67 = 1.34 \text{ watts.}$$

The portion dissipated in the load is:

$$P = 0.67^2 \times 2 = 0.9 \text{ watt,}$$

and the remainder, 0.44 watt, is dissipated in the source. If R_L is 0.5 ohm, the current in the circuit is:

$$I = \frac{2}{1 + 0.5} = 1.33 \text{ amperes.}$$

The total power is:

$$P = 2 \times 1.33 = 2.66 \text{ watts.}$$

The load dissipation is:

$$P = 1.33^2 \times 0.5 = 0.88 \text{ watt,}$$

while 1.78 watts are dissipated in the source. Thus, it is seen that, while the total dissipated power may be greater under other conditions, the dissipation in the load is greatest when its resistance equals that of the source.

Electromagnetism

The common bar or horseshoe magnet is familiar to most people. The magnetic field which surrounds it causes the magnet to attract other magnetic materials, such as iron nails or tacks. Exactly the same kind of magnetic field is set up around any conductor carrying a current, but the field exists only while the current is flowing. Such a field is called an *electromagnetic* field to distinguish it from the permanent field surrounding the bar magnet.

Magnetic Fields. Before a potential, or voltage, is applied to a conductor there is no external field, because there is no general movement of the electrons in one direction. However, the electrons do progressively move along the conductor when an e.m.f. is applied, the direction of motion depending upon the polarity of the e.m.f. Since each electron has an electric field about it, the flow of electrons causes these fields to build up into an external field which acts in a plane at right angles to the direction in which the current is flowing. This field is known as the *magnetic field.*

The magnetic field around a current-carrying conductor is illustrated in Figure 11. The direction of this magnetic field depends entirely upon the direction of electron drift or current flow in the conductor. When the flow is toward the observer, the field about the conductor is clockwise; when the flow is away from the observer, the field is counter-clockwise. This is easily remembered if the left hand is clenched, with the thumb outstretched and pointing in the direction of current flow. The fingers then indicate the direction of the magnetic field around the conductor.

Each electron adds its field to the total external magnetic field, so that the greater the number of electrons moving along the conductor, the stronger will be the resulting field.

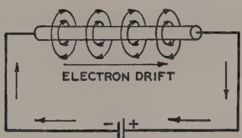


Figure 11.
Magnetic lines of force produced around a conductor carrying an electric current.

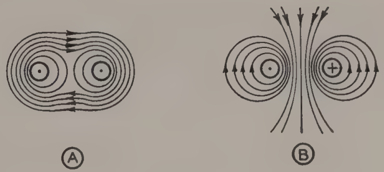


Figure 12.
Effect of (A) aiding and (B) opposing fields around adjacent conductors. Dot indicates current flowing towards observer; cross indicates current flowing away from observer.

One of the fundamental laws of magnetism is that *like poles repel one another and unlike poles attract one another.* This is true of current-carrying conductors as well as of permanent magnets. Thus, if two conductors are placed side by side and the current in each is flowing in the same direction, the magnetic fields will also be in the same direction and will combine to form a larger and stronger field, as shown in Figure 12A. If the current flow in adjacent conductors is in opposite directions, the magnetic fields will oppose each other and will tend to cancel. The magnetic field around a conductor may be considerably increased in strength by winding the wire into a coil. The field around each wire then combines with those of the adjacent turns to form a total field through the coil which is concentrated along the axis of the coil and behaves externally in a way similar to the field of a bar magnet.

If the left hand is held so that the thumb is outstretched and parallel to the axis of a coil, with the fingers curled to indicate the direction of current flow around the turns of the coil, the thumb then points in the direction of the north pole of the magnetic field.

The Magnetic Circuit. In the magnetic circuit, the units which correspond to current, voltage, and resistance in the electrical circuit are flux, magnetomotive force, and reluctance.

Flux; Flux Density. As a current is made up of a drift of electrons, so is a magnetic field made up of lines of force, and the total number of lines of force in a given magnetic circuit is termed the *flux.* The unit of flux is the *maxwell,* and the symbol is the Greek letter ϕ (phi). *Flux density* is the number of lines of force per unit area. It is expressed in *gauss* if the unit of area is the square centimeter (1 gauss = 1 line of force per square centimeter), or in *lines per square inch.* The flux depends upon the material, cross section, and length of the magnetic circuit, and it varies directly as the current flowing in the circuit.

Magnetomotive Force. The force which produces a flux in a magnetic circuit is called *magnetomotive force.* It is abbreviated m.m.f. and is designated by the letter F. The unit of magnetomotive force is the *gilbert,* which is equivalent to $1.26 \times$ NI, where N is the number of turns and I is the current flowing in the circuit in amperes.

Reluctance. Magnetic reluctance corresponds to electrical resistance, and is the property of a material that opposes the creation of a magnetic flux in the material. It is expressed in *oersteds* or in *rels,* and the symbol is the letter R. An oersted is the reluctance of 1 cubic centimeter of air. A material has a reluctance of 1 rel when an m.m.f. of 1 ampere-turn (NI) generates a flux of 1 line of force in it. Combinations of reluctances are treated the same as resistances in finding the total effective reluctance. The *specific reluctance* of any substance is its reluctance per unit volume. Excepting iron and steel and their alloys, the specific reluctance of all materials including air is 1 oersted or 0.313 rel, depending upon whether the volume is measured in cubic centimeters or in cubic inches. The symbol for specific reluctance is ν, the Greek letter nu.

As does electrical resistance, so reluctance varies directly with the length of the circuit and the specific reluctance of the material, and inversely with the area of cross section.

Ohm's Law for Magnetic Circuits. The relations between flux, magnetomotive force, and reluctance are exactly the same as the relations between current, voltage, and resistance in the electrical circuit. These can be stated as follows:

$$\phi = \frac{F}{R} \qquad R = \frac{F}{\phi} \qquad F = \phi R$$

where ϕ = flux, F = m.m.f., and R = reluctance. If F is in gilberts, R will be expressed in oersteds, but if F is in ampere-turns, then R will be in rels.

Calculations. To simplify magnetic circuit calculations, a magnetization curve may be drawn for a given unit of material. Such a curve is termed a B-H curve, and is arrived at by experiment. B-H curves for most common

magnetic materials are available in many reference books, so none will be given here.

The symbol for flux density is B if it is expressed in gausses, or B if expressed in lines per square inch. The m.m.f. necessary to produce a given flux density is stated in gilberts per centimeter (H), or in ampere-turns per inch (H).

As an example, a transformer is to be designed with a core of open-hearth steel having a cross-sectional area of 3 sq. cm. The flux is to be 36,000 lines. The magnetic path is 24 cm. long. It is necessary to find the ampere-turns required to produce the desired flux. $B = \phi/\mathrm{A} = 36{,}000/3 = 12{,}000$ gausses or 12 kilogausses. A B-H curve for open-hearth steel shows that 5 gilberts per centimeter are required to magnetize this material to a flux density of 12 kilogausses. Since the path is 24 cm. long, 5 must be multiplied by 24, giving 120 gilberts required. To obtain ampere-turns, 120 is divided by 1.26, giving 95.2 ampere-turns.

Permeability. Permeability expresses the ease with which a magnetic field may be set up in a material as compared with the effort required in the case of air. Iron, for example, has a permeability of around 2000 times that of air, which means that a given amount of magnetizing effect produced in an iron core by a current flowing through a coil of wire will produce 2000 times the *flux density* that the same magnetizing effect would produce in air. It may be expressed by the ratio B/H or B/H. In other words,

$$\mu = \frac{B}{H} \text{ or } \mu = \frac{\mathrm{B}}{\mathrm{H}}$$

where μ is the permeability, B is the flux density in gausses, B is the flux density in lines per square inch, H is the m.m.f. in gilberts per centimeter, and H is the m.m.f. in ampere-turns per inch. These relations may also be stated as follows:

$$H = \frac{B}{\mu} \text{ or } \mathrm{H} = \frac{\mathrm{B}}{\mu}, \text{ and } B = H\mu \text{ or } \mathrm{B} = \mathrm{H}\mu$$

Saturation. Permeability is similar to *electric conductivity*. There is, however, one important difference: the permeability of magnetic materials is not independent of the magnetic current (flux) flowing through it, although electrical conductivity is usually independent of the electric current in a wire. When the flux density of a magnetic conductor has been increased to the *saturation point*, a further increase in the magnetizing force will not produce a material increase in flux density.

Residual Magnetism; Retentivity. The magnetism remaining in a material after the magnetizing force is removed is called *residual magnetism*. *Retentivity* is the property which causes a magnetic material to have residual magnetism after having been magnetized.

Hysteresis; Coercive Force. *Hysteresis* is the characteristic of a magnetic system which causes a loss of power due to the fact that a negative magnetizing force must be applied to reduce the residual magnetism to zero. This negative force is termed *coercive force*. By "negative" magnetizing force is meant one which is of the opposite polarity with respect to the original magnetizing force. Hysteresis loss is apparent in transformers and chokes by the heating of the core.

Alternating Current

To this point in the text, consideration has been given primarily to a current consisting of a steady flow of electrons in one direction. This type of current flow is known as *unidirectional* or *direct* current, abbreviated *d.c.* Equally as important in radio work and more important in power practice is another and altogether different type of current, known as *alternating current* and abbreviated *a.c.* Power distribution from one point to another and into homes and factories is almost universally a.c. On the other hand, the plate supply to vacuum tubes is almost universally d.c.

Generation of Alternating Current. Faraday discovered that, if a conductor which forms part of a closed circuit, is moved through a magnetic field so as to cut across the lines of force, a current will flow in the conductor. He also discovered that, if a conductor in a second closed circuit is brought near the first conductor and the current in the first one is varied, a current will flow in the second conductor. This effect is known as *induction*, and the currents so generated are *induced currents*. In the latter case it is the lines of force which are moving and cutting the second conductor, due to the varying current strength in the first conductor.

A current is induced in a conductor if there is a relative motion between the conductor and a magnetic field, its direction of flow depending upon the direction of the relative motion between the conductor and the field, and its strength depends upon the intensity of the field, the rate of cutting lines of force, and the number of turns in the conductor.

An alternating current is one which periodically rises from zero to a maximum in one direction, decreases to zero and changes its direction, rises to a maximum in the opposite direction, and decreases to zero again. This

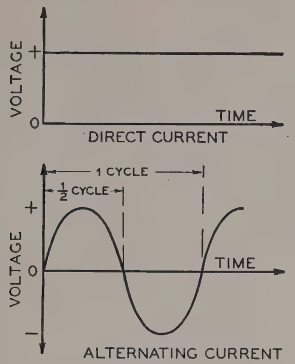


Figure 13.
Graphical comparison of unidirectional or direct current and alternating current.

complete process is called a *cycle,* and from zero through a maximum and back to zero is an *alternation* or *half-cycle.* The number of times per second that the current goes through a complete cycle is called the *frequency.*

A machine that generates alternating current is termed an *alternator* or *a.c. generator.* Such a machine in its basic form is shown in Figure 14. It consists of two permanent magnets, M, the opposite poles of which face each other, and the poles are machined so that they have a common radius. Between these two poles, *north* (N) and *south* (S), a magnetic field exists. If a conductor in the form of C is so suspended that it can be freely rotated between the two poles, and if the opposite ends of conductor C are brought to collector rings, R, which are contacted by brushes (B), there will be a flow of alternating current when conductor C is rotated. This current will flow out through the collector rings R and brushes B to the external circuit, X-Y.

The field intensity between the two pole pieces is substantially constant over the entire area of the pole face. However, when the conductor is moving parallel to the lines of force at the top or bottom of the pole faces, no lines are being cut. As the conductor moves on across the pole face it cuts more and more lines of force for each unit distance of travel, until it is cutting the maximum number of lines when opposite the center of the pole. Therefore, zero current is induced in the conductor when it is at the top or bottom of the poles, and maximum current is induced when it is opposite the center of the pole face. After the conductor has rotated through 180° it can be seen that its position with respect to the pole pieces will be exactly opposite to that when it started. Hence, the second 180° of rotation will produce an alternation of current in the opposite direction to that of the first alternation.

The current does *not* increase directly as the angle of rotation, but rather as the *sine* of the angle; hence, such a current has the mathematical form of a *sine wave.* Although most electrical machinery does not produce a strictly pure sine curve, the departures are usually so slight that the assumption can be regarded as fact for most practical purposes. All that has been said in the foregoing paragraphs concerning alternating current also is applicable to alternating voltage.

Referring to Figure 15, it will be seen that if a curve is plotted for an alternating voltage, such a curve would assume the shape of a sine wave and by plotting amplitude against time, the voltage at any instant could be found. The instantaneous value of voltage at any given instant can also be calculated as follows:

$$e = E_{max} \sin 2\pi ft,$$

where e = the instantaneous voltage,
E = maximum crest value of voltage,
f = frequency in cycles per second, and
t = time in seconds.

The instantaneous current can be found from the same formula by substituting i for e and I_{max} for E_{max}. The formula then becomes:

$$i = I_{max} \sin 2\pi ft,$$

where i = the instantaneous current,
I = maximum crest value of current,
f = frequency in cycles per second, and
t = time in seconds.

Radians. The term $2\pi f$ in the preceding equation should be thoroughly understood because it is of basic importance. Returning again to the rotating point P (Figure 15), it can be seen that when this point leaves its horizontal position and begins its rotation in a counter-clockwise direction, through a complete revolution back to its initial starting

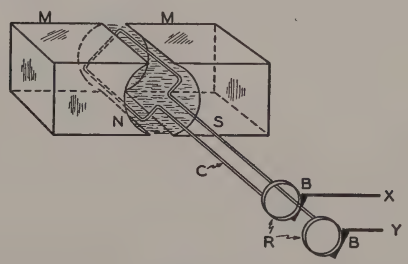


Figure 14.
Schematic representation of the simplest form of the alternator.

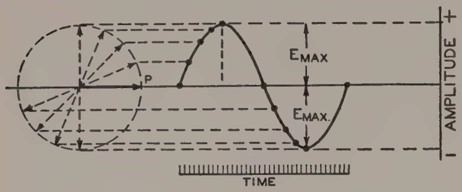


Figure 15.
Graph showing the voltage output of a single-turn conductor revolving in a magnetic field.

point, it will have traveled through 360 electrical degrees. In electrical work, instead of referring to this movement in terms of degrees, it is customary to express the movement in terms of *radians.* Mathematically, a radian is an arc of the circle equal in length to the radius of the circle. There are 2π radians in 360 degrees, so that one radian is equivalent to 57.32 degrees.

When the conductor in the simple alternator has moved through 2π radians it has generated one cycle. $2\pi f$ then represents one cycle, multiplied by the number of cycles per second (the frequency) of the alternating voltage or current, and is, therefore, the *angular velocity.* In technical literature $2\pi f$ is often replaced by ω, the Greek letter omega. Velocity multiplied by time gives the distance traveled, so $2\pi ft$ represents the *angular distance* through which the conductor has traveled, and since the instantaneous voltage or current is proportional to the sine of this angle, it is possible to calculate these quantities at any instant of time, provided that the wave very closely approximates a sine curve.

Vectors; the J Operator. The quantities mentioned so far have been *scalar;* that is, they have only a magnitude and can be expressed completely by a number. However, a.c. problems can often be simplified if use is made of *vectors,* which indicate the direction of a quantity as well as its magnitude. The idea of a quantity having direction may be somewhat confusing, but it will be made clear in the following discussion.

In a.c. work two types of vectors may occur, *space* and *rotating* vectors. A space vector is one which has a fixed direction in space, while a rotating vector has a constant magnitude and angular velocity. Alternating currents or voltages which vary sinusoidally with time are usually expressed by rotating vectors. The projection of the vector upon a fixed reference axis represents the instantaneous value of the current or voltage, if the length of the vector indicates the maximum value.

Vector quantities may be written in a form known as *complex notation.* To do this requires the use of the *j* operator. It is possible to determine the position of the hypotenuse of a right triangle, with respect to a system of rectangular coordinate axes, if the length and mathematical sign of the two sides adjacent to the right angle are known. Referring to Figure 18, if the lengths of OA and AP in the triangle AOP are known, the location of the hypotenuse OP is fixed, since both of the sides are positive. By using *j* to indicate that AP (or OC) is at right angles to OA, OP can be written as OA+*j*AP. Depending upon whether the side to which it refers extends upward or downward from the horizontal axis, *j* can be either positive or negative. Likewise, OA can be either positive or negative, depending upon whether it extends to the right or left of the vertical axis.

Frequency. The frequency of an alternating current of voltage may be any value from 1 cycle per second to an infinitely large number of cycles per second. Up to about 20,000 cycles per second are considered audio frequencies, since all except those from zero to about 16 c.p.s. are audible to the human ear. The a.c. power which is supplied to homes and factories is generally 25, 50, or 60 c.p.s. Frequencies above 20,000 c.p.s. are known as radio frequencies. But they are usually spoken of in terms of kilocycles, rather than cycles, because the numbers become too large. When the frequency gets above a few thousand kilocycles, the term megacycles is used. A kilocycle is equal to 1000 cycles, and a megacycle equals 1,000,000 cycles. A conversion table for simplifying this terminology is given here:

1,000 *cycles* = 1 *kilocycle. The abbreviation for kilocycle is kc.*
1 *cycle* = 1/1,000 *of a kilocycle,* .001 *kc. or* 10^{-3} *kc.*
1 *megacycle* = 1,000 *kilocycles, or* 1,000,000 *cycles,* 10^3 *kc. or* 10^6 *cycles.*

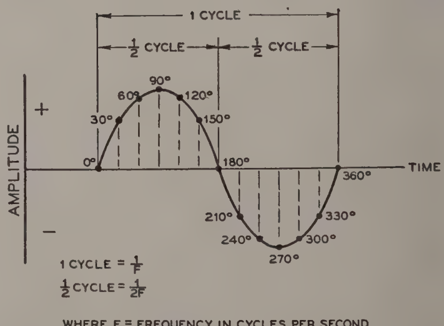


Figure 16.

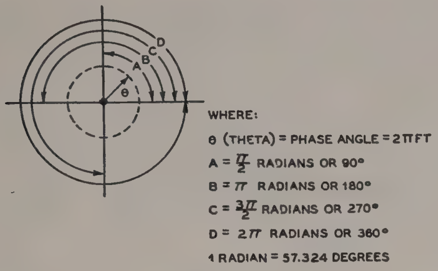


Figure 17.

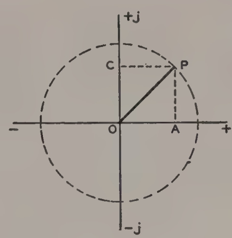


Figure 18.
ROTATING VECTOR DIAGRAM.

1 *kilocycle* = 1/1,000 *megacycle,* .001 *megacycle, or* 10^{-3} *Mc. The abbreviation for megacycles is Mc.*

Effective Value of Voltage and Current. The instantaneous value of an alternating current or voltage varies throughout the cycle, so that the *effective* value of this current or voltage must be determined by comparing the a.c. heating effect with that of d.c. Thus, an alternating current will have an effective value of 1 ampere when it produces the same heat in a conductor as does 1 ampere of direct current.

This effective value is derived by taking the instantaneous values of current over a cycle of alternating current, squaring these values, taking an average of the squares, and then taking the square root of the average. By this procedure, the effective value becomes known as the *root mean square* or r.m.s. value. This is the value that is read on a.c. voltmeters and ammeters. The r.m.s. value is 70.7 per cent of the peak or maximum instantaneous value and is expressed as follows:

$$E_{eff} \text{ or } E_{r.m.s.} = 0.707 \times E_{max}, \text{ or}$$
$$I_{eff} \text{ or } I_{r.m.s.} = 0.707 \times I_{max}.$$

The following relations are extremely useful in radio and power work:

$$E_{r.m.s.} = 0.707 \times E_{max}, \text{ and}$$
$$E_{max} = 1.414 \times E_{r.m.s.}$$

Rectified Alternating Current or Pulsating Direct Current. If an alternating current is passed through a full-wave rectifier, it emerges in the form of a current of *varying amplitude* which flows in *one* direction only. Such a current is known as *rectified a.c.* or *pulsating d.c.* A typical wave form of a pulsating direct current as would be obtained from the output of a full-wave rectifier is shown in Figure 19.

Measuring instruments designed for d.c. operation will not read the peak or instantaneous maximum value of the pulsating d.c. output from the rectifier; it will read only the *average value*. This can be explained by assuming that it could be possible to cut off some of the peaks of the waves, using the cut-off portions to fill in the spaces that are open, thereby obtaining an *average* d.c. value. A milliammeter and voltmeter connected to the adjoining circuit, or across the output of the rectifier, will read this average value. It is related to *peak* value by the following expression:

$$E_{avg} = 0.636 \times E_{max}$$

It is thus seen that the average value is 63.6 per cent of the peak value.

Relationship Between Peak, R.M.S. or Effective, and Average Values. To summarize the three most significant values of an a.c. wave: the peak value is equal to 1.41 times the r.m.s. or effective, and the r.m.s. value is equal to 0.707 times the peak value; the average value of a full-wave rectified a.c. wave is 0.636 times the peak value, and the average value of a rectified wave is equal to 0.9 times the r.m.s. value. This latter factor is of value in determining the voltage output from a power supply which operates with a choke-input filter system. If the input choke is of the swinging type and is of ample inductance, the d.c. voltage output of the power supply will be 0.9 times the r.m.s. a.c. output of the used secondary of the transformer (one-half secondary voltage in the case of a full-wave rectifier and the full secondary voltage in the case of bridge rectification) less the drop in the rectifier tubes (usually negligible) and the drop in the filter inductances.

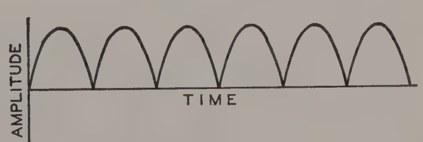
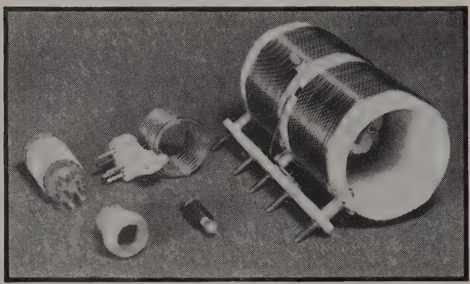


Figure 19.
Typical waveform output as obtained from a full-wave rectifier.

Inductance

In the section titled *"Generation of Alternating Current"* a brief explanation of induction was given, and it would be well for the reader to review it at this point.

If a switch is inserted in the circuit shown in Figure 11, a pulsating direct current can be produced by closing and opening the switch. When it is first closed, the current does not instantaneously rise to its maximum value, but builds up gradually. While it is building up the magnetic field is expanding around the conductor. Of course, this happens in a fraction of a second. If the switch is then opened, the current dies down gradually and the magnetic field contracts. This expanding and contracting field will induce a current in any other conductor that is part of a continuous circuit which it cuts. Such a field can be obtained in the way just mentioned, by means of a vibrator interruptor, or by applying a.c. to the circuit in place of the battery. Varying



the resistance of the circuit will also produce the same effect. This inducing of a current in a conductor due to a varying current in another conductor is called *electromagnetic induction.*

Self-Induction. If an alternating current flows through a coil the varying magnetic field around each turn cuts itself and the adjacent turns and induces a voltage in the coil, of opposite polarity to the applied e.m.f. The amount of induced voltage depends upon the number of turns in the coil, the current flowing in the coil, and the number of lines of force threading the coil. The voltage so induced is known as a *counter-e.m.f.* or *back-e.m.f.*, and the effect is termed *self-induction.* When the applied voltage is building up the counter-e.m.f. opposes the rise; when the applied voltage is decreasing the counter-e.m.f. is of the same polarity and tends to maintain the current. Thus, it can be seen that self-induction tends to prevent any change in the current in the circuit.

The Unit of Inductance; The Henry. Inductance is usually denoted by the letter L, and is expressed in *henrys.* A coil has an inductance of 1 henry when a voltage of 1 volt is induced by a current change of 1 ampere per second. The henry, while commonly used in audio frequency circuits, is too large for reference to inductance coils such as those used in radio frequency circuits; *millihenry* or *microhenry* are more commonly used, in the following manner:

1 *henry* = 1,000 *millihenrys, or* 10^3 *millihenrys.*

1 *millihenry* = 1/1,000 *of a henry*, .001 *henry, or* 10^{-3} *henry.*

1 *microhenry* = 1/1,000,000 *of a henry, or* .000001 *henry, or* 10^{-6} *henry.*

One one-thousandth of a millihenry = .001 *or* 10^{-3} *millihenrys.*

1,000 *microhenrys* = 1 *millihenry.*

Mutual Induction. When one coil is near another, a varying current in one will produce a varying magnetic field which cuts the turns of the other coil, inducing a current in it. This induced current is also varying, and will therefore induce another current in the first coil. This reaction between two coupled circuits is called *mutual induction,* and can be calculated and expressed in henrys. The symbol for mutual inductance is M. Two circuits thus joined are said to be *inductively coupled.*

The magnitude of the mutual inductance depends upon the shape and size of the two circuits, their positions and distances apart, and the permeability of the medium. The extent to which two inductances are coupled is expressed by a relation known as *coefficient of coupling.* This is the ratio of the mutual inductance actually present to the maximum possible value.

Inductances in Parallel. Inductances in parallel are combined exactly as are resistors in parallel, provided that they are far enough apart so that the mutual inductance is entirely negligible, i.e., if the coupling is very loose.

Inductances in Series. Inductances in series are additive, just as are resistors in series, again provided that no mutual inductance exists. In this case, the total inductance L is:

$$L = L_1 + L_2 + \ldots\ldots\ldots\ldots \text{etc.}$$

Where mutual inductance does exist:

$$L = L_1 + L_2 + 2M,$$

where M is the mutual inductance.

This latter expression assumes that the coils are connected in such a way that all flux linkages are in the same direction, i.e., additive. If this is not the case and the mutual linkages

subtract from the self-linkages, the following formula holds:

$$L = L_1 + L_2 - 2M,$$

where M is the mutual inductance.

Calculation of Inductance. The inductance of coils with magnetic cores can be determined with reasonable accuracy from the formula:

$$L = 1.257 \times N^2 \times P \times 10^{-3}$$

where

L is the inductance in henrys,
N is the number of turns,
P is the permeability of the core material.

From this formula it can be seen that the inductance is proportional to the *permeability* as well as to the square of the number of turns. Thus, it is possible to secure greater values of inductance with a given number of turns of wire wound on an iron core than would be possible if an air core coil were used.

The inductance of an air core coil can be calculated with good accuracy by using the following formula:

$$L = \frac{0.2 \times D^2 \times N^2}{3D + 9l + 10C}$$

where L = inductance in microhenrys,
D = inside diameter of coil in inches,
l = length of winding in inches,
C = radial depth of coil in inches (this value may be omitted for single-layer coils), and
N = total number of turns.

By first deciding upon the size of wire that is to be used, l can be expressed in terms of N as follows:

$$l \text{ (length of winding)} = \frac{\text{total number of turns in coil}}{\text{number of turns per linear inch}}, \text{ or } l = \frac{K}{N}.$$

Core Material. Ordinary magnetic cores cannot be used for radio frequencies because the *eddy current losses* in the core material become enormous as the frequency is increased. The principal use for magnetic cores is in the audio-frequency range below approximately 15,000 cycles, whereas at very low frequencies (50 to 60 cycles) their use is mandatory if an appreciable value of inductance is desired.

An air core inductor of only 1 henry inductance would be quite large in size, yet values as high as 500 henrys are commonly available in small iron core chokes. The inductance of a coil with a magnetic core will vary with the amount of direct current which passes through the coil. For this reason, iron core chokes that are used in power supplies have a certain inductance rating at a *predetermined value of d.c.*

One exception to the statement that metal core inductances are highly inefficient at radio frequencies is in the *powdered* iron cores used in some types of intermediate frequency transformers. These cores are made of very *fine particles* of powdered iron, which are first treated with an insulating compound so that each particle is insulated from the other. These particles are then molded into a solid core around which the wire is wound. Eddy current losses are greatly reduced, with the result that these special iron cores are entirely practical in circuits which operate up to 1500 kc. in frequency.

Inductive Reactance. As was previously stated, when an alternating current flows through an inductance, a back- or counter-electromotive force is developed; this force opposes any change in the initial e.m.f. The property of an inductance to offer opposition to a change in current is known as its *reactance* or *inductive reactance.* This is expressed as X_L:

$$X_L = 2\pi fL,$$

where X_L = inductive reactance expressed in ohms.

$\pi = 3.1416$ ($2\pi = 6.283$),
f = frequency in cycles,
L = inductance in henrys.

Inductive Reactance at R. F. It is very often necessary to compute inductive reactance at radio frequencies. The same formula may be used, except that the units in which the inductance and the frequency are expressed will be changed. Inductance can, therefore, be expressed in *millihenrys* and frequency in *kilocycles.* For higher frequencies and smaller values of inductance, frequency is expressed in *megacycles* and inductance in *microhenrys.* The basic equation need not be changed, since the multiplying factors for inductance and frequency appear in numerator and denominator, and hence are cancelled out. However, it is not possible in the same equation to express L in millihenrys and f in cycles without conversion factors.

Should it become desirable to know the value of inductance necessary to give a certain reactance at some definite frequency, a transposition of the original formula gives the following:

$$L = X_L \div (2\pi f),$$

or when X_L and L are known,

$$f = \frac{X_L}{2 \pi L}.$$

Electrostatic Storage of Energy

So far we have dealt only with the storage of energy in an electromagnetic field in the form of an inductance. The storage of energy in a magnetic field is expressed in *joules* and is equal to $(LI^2)/2$. (A joule is equal to 1 watt-second.) Electrical energy can also be stored in an electrostatic field. A device capable of storing energy in such a field is called a *condenser* and is said to have a certain *capacitance*. The energy stored in an electrostatic field is also expressed in *joules* and is equal to $CE^2/2$, where C is the capacity in *farads* (a unit of capacity to be discussed) and E is the potential in volts.

Capacitance and Condensers. Two metallic plates separated from each other by a thin layer of insulating material (called a *dielectric,* in this case), become a *condenser.* When a source of d.c. potential is momentarily applied across these plates, they may be said to become charged. If the same two plates are then joined together momentarily by means of a wire, the condenser will *discharge*.

When the potential was first applied, electrons immediately attempted to flow from one plate to the other through the battery or such source of d.c. potential as was applied to the condenser plates. However, the circuit from plate to plate in the condenser was *incomplete* (the two plates being separated by an insulator) and thus the electron flow ceased, meanwhile establishing a shortage of electrons on one plate and a surplus of electrons on the other.

Remember that when a deficiency of electrons exists at one end of a conductor, there is always a tendency for the electrons to move about in such a manner as to re-establish a state of balance. In the case of the condenser herein discussed, the surplus quantity of electrons on one of the condenser plates cannot move to the other plate because the circuit has been broken; that is, the battery or d.c. potential was removed. This leaves the condenser in a *charged* condition; the condenser plate with the electron *deficiency* is *positively* charged, the other plate being *negative*.

In this condition, a considerable stress exists in the insulating material (dielectric) which separates the two condenser plates, due to the mutual attraction of two unlike potentials on the plates. This stress is known as *electrostatic* energy, as contrasted with *electromagnetic* energy in the case of an inductance. This charge can also be called *potential energy* because it is capable of performing work when the charge is released through an external circuit.

The Unit of Capacitance: The Farad. If the external circuit of the two condenser plates is completed by joining the terminals together with a piece of wire, the electrons will rush immediately from one plate to the other through the external circuit and establish a state of equilibrium. This latter phenomenon explains the *discharge* of a condenser. The amount of stored energy in a charged condenser is dependent upon the charging potential, as well as a factor which takes into account the *size* of the plates, *dielectric thickness, nature* of the dielectric, and the *number* of plates. This factor, which is determined by the foregoing, is called the *capacity* of a condenser and is expressed in *farads.*

The farad is such a large unit of capacity that it is rarely used in radio calculations, and the following more practical units have, therefore, been chosen:

1 *microfarad* = 1/1,000,000 *of a farad, or* .000001 *farad, or* 10^{-6} *farads.*

1 *micro-microfarad* = 1/1,000,000 *of a microfarad, or* .000001 *microfarad, or* 10^{-6} *microfarads.*

1 *micro-microfarad = one-million-millionth of a farad, or* .000000000001 *farad, or* 10^{-12} *farads.*

If the capacity is to be expressed in *microfarads* in the equation given under *energy storage,* the factor C would then have to be divided by 1,000,000, thus:

$$\text{Stored energy in joules} = \frac{C \times E^2}{2 \times 1{,}000{,}000}$$

This storage of energy in a condenser is one of its very important properties, particularly in those condensers which are used in power supply filter circuits.

Dielectric Constant. The capacity of a condenser is largely determined by the thickness and nature of the dielectric separation between plates. Certain materials offer a greater capacity than others, depending upon their physical makeup and chemical constitution. This property is expressed by a constant K, called the dielectric constant. A table for

some of the commonly used dielectrics is given here:

Material	*Dielectric Constant*
Air	1.00
Mica	5.75
Hard rubber	2.50 to 3.00
Glass	4.90 to 9.00
Bakelite derivatives	3.50 to 6.00
Celluloid	4.10
Fiber	4 to 6
Wood (without special preparation):	
Oak	3.3
Maple	4.4
Birch	5.2
Transformer oil	2.5
Castor oil	5.0
Porcelain, steatite	6.5
Lucite	2.5 to 3.0
Quartz	4.75
Victron, Trolitul	2.6

Dielectric Breakdown. The nature and thickness of a dielectric have a very definite bearing on the amount of charge of a condenser. If the charge becomes too great for a given thickness of dielectric, the condenser will break down, i.e., the dielectric will puncture. It is for this reason that condensers are rated in the manner of the amount of voltage they will safely withstand. This rating is commonly expressed as the *d.c. working voltage.*

Calculation of Capacity. The capacity of two parallel plates is given with good accuracy by the following formula:

$$C = 0.2248 \times K \times \frac{A}{t},$$

where C = capacity in micro-microfarads,
K = dielectric constant of spacing material,
A = area of dielectric in square inches,
t = thickness of dielectric in inches.

This formula indicates that the capacity is directly proportional to the area of the plates and inversely proportional to the thickness of the dielectric (spacing between the plates). This simply means that when the area of the plate is doubled, the spacing between plates remaining constant, the capacity will be doubled. Also, if the area of the plates remains constant, and the plate spacing is doubled, the capacity will be reduced to half. The above equation also shows that capacity is directly proportional to the dielectric constant of the spacing material. A condenser that has a capacity of 100 $\mu\mu$fd. in air would have a capacity of 500 $\mu\mu$fd. when immersed in castor oil, because the dielectric constant of castor oil is

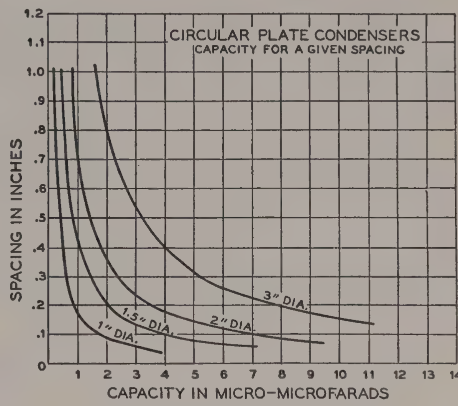


Figure 20.

5.0 or five times as great as the dielectric constant of air.

Where the area of the plates is definitely set, and when it is desired to know the spacing needed to secure a required capacity,

$$t = \frac{A \times 0.2248 \times K}{C},$$

where all units are expressed just as in the preceding formula. This formula is not confined to condensers having only square or rectangular plates, but also applies when the plates are circular in shape. The only change will be the calculation of the *area* of such circular plates; this area can be computed by squaring the *radius* of the plate, then multiplying by 3.1416, or "pi." Expressed as an equation:

$$A = 3.1416 \times r^2,$$

where r = radius in inches.

The capacity of a multi-plate condenser can be calculated by taking the capacity of one section and multiplying this by the number of dielectric spaces. In such cases, however, the formula gives no consideration to the effects of edge capacity so that the capacity as calculated will not be entirely accurate. These additional capacities will be but a small part of the effective total capacity, particularly when the plates are reasonably large, and the final result will, therefore, be within practical limits of accuracy.

Equations for calculating capacities of condensers in *parallel* connection are the same as those for resistors in *series:*

$$C = C_1 + C_2, \text{ etc.}$$

Condensers in *series* connection are calculated in the same manner as are resistors in *parallel* connection.

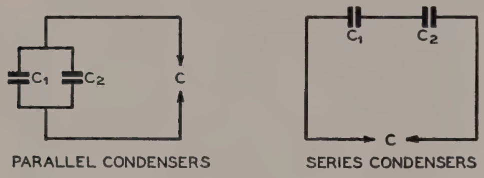


Figure 21.

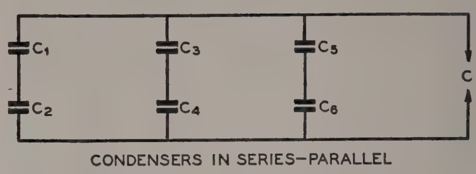


Figure 22.

The formulas are repeated: (1) For two or more condensers of *unequal* capacity in series:

$$C = \frac{1}{\frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}},$$

$$\text{or } \frac{1}{C} = \frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}$$

(2) *Two* condensers of *unequal* capacity in series:

$$C = \frac{C_1 \times C_2}{C_1 + C_2}$$

(3) Three condensers of *equal* capacity in series:

$C = \frac{C_1}{3}$, where C_1 is the common capacity.

(4) Three or more condensers of *equal* capacity in series:

$$C = \frac{\text{Value of common capacity}}{\text{Number of condensers in series}}$$

(5) Six condensers in series parallel:

$$C = \frac{1}{\frac{1}{C_1} + \frac{1}{C_2}} + \frac{1}{\frac{1}{C_3} + \frac{1}{C_4}} + \frac{1}{\frac{1}{C_5} + \frac{1}{C_6}}$$

Capacitive Reactance. It has been explained that inductive reactance is the ability of an inductance to oppose a change in an alternating current. Condensers have a similar property although in this case the opposition is to the *voltage* which acts to charge the condenser. This action is called *capacitive reactance* and is expressed as follows:

$$X_c = \frac{1}{2\pi fC},$$

where X_c = capacitive reactance in ohms,
π = 3.1416,
f = frequency in cycles,
C = capacity in farads.

Capacitive Reactance at R. F. Here again, as in the case of inductive reactance, the units of capacity and frequency can be converted into smaller units for practical problems encountered in radio work. The equation may be written:

$$X_c = \frac{1{,}000{,}000}{2\pi fC},$$

where f = frequency in megacycles,
C = capacity in micro-microfarads.

In the design of filter circuits, it is often convenient to express frequency (f) in *cycles* and capacity (C) in *microfarads,* in which event the same formula applies.

Condensers in A. C. and D. C. Circuits. When a condenser is connected into a direct current circuit, it will block the d.c., or stop the flow of current. Beyond the initial movement of electrons during the period when the condenser is being charged, there will be no flow of current because the circuit is effectively broken by the dielectric of the condenser. Strictly speaking, a very small current may actually flow because the dielectric of the condenser may not be a perfect insulator. This minute current flow is the leakage current previously referred to and is dependent upon the internal d.c. resistance of the condenser. This leakage current is usually quite noticeable in most types of electrolytic condensers.

When an alternating current is applied to a condenser, the condenser will charge and discharge a certain number of times per second in accordance with the frequency of the alternating voltage. The electron flow in the charge and discharge of a condenser when an a.c. potential is applied constitutes an alternating current, in effect. It is for this reason that a condenser will pass an alternating current yet offer practically infinite opposition to a direct current. These two properties are repeatedly in evidence in a radio circuit.

Voltage Rating of Condensers in Series. Any good paper dielectric filter condenser has such a high internal resistance (indicating a good dielectric) that the exact resistance will vary considerably from condenser to condenser even though they are made by the same manufacturer and are of the same rating. Thus,

when 1000 volts d.c. is connected across two 1-μfd. 500-volt condensers, the chances are that the voltage will divide unevenly and one condenser will receive more than 500 volts and the other less than 500 volts.

Voltage Equalizing Resistors. By connecting a half-megohm 1-watt carbon resistor across each condenser, the voltage will be equalized because the resistors act as a voltage divider, and the internal resistances of the condensers are so much higher (many megohms) that they have but little effect in disturbing the voltage divider balance.

Carbon resistors of the inexpensive type are not particularly accurate (not being designed for precision service); therefore it is advisable to check several on an accurate ohmmeter to find two that are as close as possible in resistance. The exact resistance is unimportant, just so it is the same for the two resistors used.

Condensers in Series on A. C. When two condensers are connected in series, *alternating* current pays no heed to the relatively high internal resistance of each condenser, but divides across the condensers in inverse proportion to the *capacity*. Because, in addition to the d.c. across a capacitor in a filter or audio amplifier circuit, there is usually an a.c. or a.f. voltage component, it is inadvisable to series-connect condensers of unequal capacitance even if dividers are provided to keep the d.c. within the ratings of the individual capacitors.

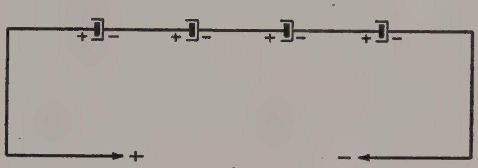
For instance, if a 500-volt 1-μfd. capacitor is used in series with a 4-μfd. 500-volt condenser across a 250-volt a.c. supply, the 1-μfd. condenser will have 200 volts a.c. across it and the 4-μfd. condenser only 50 volts. An equalizing divider to do any good in this case would have to be of very low resistance because of the comparatively low impedance of the condensers *to a.c.* Such a divider would draw excessive current and be impracticable.

The safest rule to follow is to use only condensers of the same capacity and voltage rating and to install matched high resistance proportioning resistors across the various condensers to equalize the d.c. voltage drop across each condenser. This holds regardless of how many capacitors are series-connected.

Electrolytic Condensers in Series. Similar electrolytic capacitors, of the same capacity and made by the same manufacturer, have more nearly uniform (and much lower) internal resistance though it still will vary considerably. However, the variation is not nearly as great as encountered in paper condensers, and the lowest d.c. voltage is across the weakest (leakiest) electrolytic condensers of a series group.

As an electrolytic capacitor begins to show signs of breaking down from excessive voltage, the leakage current goes up, which tends to heat the condenser and aggravate the condition. However, when used in series with one or more others, the lower resistance (higher leakage current) tends to put less d.c. voltage on the weakening condenser and more on the remaining ones. Thus, the capacitor with the *lowest* leakage current, usually the *best* capacitor, has the highest voltage across it. For this reason, dividing resistors are not essential across series-connected electrolytic capacitors.

Electrolytic condensers use a very thin film of oxide as the dielectric, and are polarized; that is, they have a positive and a negative terminal which must be properly connected in a circuit; otherwise, the oxide will boil, and the condenser will no longer be of service. When electrolytic condensers are connected in series, the positive terminal is always connected to the positive lead of the power supply; the negative terminal of the condenser connects to the positive terminal of the *next* condenser in the series combination. The method of connection is shown in Figure 23.



POLARIZED CONDENSERS (ELECTROLYTIC) IN SERIES

Figure 23.

Phase. When an alternating current flows through a purely resistive circuit, it will be found that the current will go through maximum and minimum in perfect step with the voltage. In this case the current is said to be in step or *in phase* with the voltage. For this reason, Ohm's law will apply equally well for a.c. or for d.c. where pure resistances are concerned, provided that the *effective* values of a.c. are used in the calculations.

If a circuit has capacity or inductance or both, in addition to resistance, the current does not reach a maximum at the same instant as the voltage; therefore Ohm's law will *not* apply. It has been stated that inductance tends to resist any change in current; when an inductance is present in a circuit through which an alternating current is flowing, it will be found that the current will reach its maximum *behind* or later than the voltage. In electrical terms, the current will *lag* behind the voltage, or, conversely, the voltage will *lead* the current.

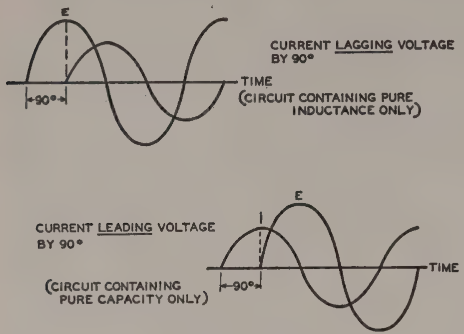


Figure 24.

The above two illustrations show the manner in which a pure inductance or a pure capacitance (no resistance component in either) will cause the current in the circuit either to lead or to lag the voltage by 90°.

If the circuit is *purely* inductive, i.e., if it contains neither resistance nor capacitance, the current does not start until the voltage has first reached a maximum; the current, therefore, *lags* the voltage by 90 degrees as in Figure 24. The angle will be less than 90 degrees if resistance is in the circuit.

When pure *capacity* alone is present in an a.c. circuit (no inductance or resistance of any kind), the opposite effect will be encountered; the current will reach a maximum at the instant the voltage is starting and, hence, will *lead* the voltage by 90 degrees. The presence of resistance in the circuit will tend to decrease this angle.

Comparison of Inductive to Capacitive Reactance with Changing Frequency. From the equation for *inductive* reactance, it is seen that as the frequency becomes greater the reactance increases in a corresponding manner. The reactance is doubled when the frequency is doubled. If the reactance is to be very large when the frequency is low, the value of inductance must be very large.

The equation for capacitive reactance shows that the reactance varies *inversely* with frequency and capacity. With a fixed value of capacity, the reactance will become less as the frequency increases. When the frequency is fixed, the reactance will be greater as the capacity is lowered. In order to have high reactance, it is necessary to have low capacitance although in power filter circuits the reactance is always made low so that the alternating current component from the rectifier will be bypassed. The capacitance must be made large in this case because the frequency is quite low (60–120 cycles).

A comparison of the two types of reactance, inductive and capacitive, shows that in one case (inductive) the reactance *increases* with frequency, whereas in the other (capacitive) the reactance *decreases* with frequency.

Reactance and Resistance in Combination. When a circuit includes a capacity or an inductance or both, in addition to a resistance, the simple calculations of Ohm's law will *not* apply when the total impedance to alternating current is to be determined. Reference is here made to the passage of an *alternating current* through the circuit; the reactance must be considered in addition to the d.c. resistance because reactance offers an opposition to the flow of alternating current.

When alternating current passes through a circuit which contains only a condenser, the voltage and current relations are as follows:

$$E = IX_c, \text{ and } I = \frac{E}{X_c},$$

where E = voltage,
I = current in amperes,
X_c = capacitive reactance or $\dfrac{1}{2\pi fC}$
(expressed in ohms).

Power Factor. It should now be apparent to the reader that in such circuits that have reactance as well as resistance, it will not be possible to calculate the power as in a d.c. circuit or as in an a.c. circuit in which current and voltage are in-phase. The reactive components cause the voltage and current to reach their maximums at different times, as was explained under *Phase,* and to calculate the power in such a circuit we must use a value called the *power factor* in our computations.

The *power factor* in a resistive-reactive a.c. circuit may be expressed as the *actual* watts (as measured by a watt-meter) divided by the product of voltage and current or:

$$\frac{W}{E \times I}$$

where W = watts as measured,
E = voltage (r.m.s.)
I = current in amperes (r.m.s.).

Stated in another manner:

$$\frac{W}{E \times I} = \cos\theta$$

The character θ is the angle of phase difference between current and voltage. The product of volts times amperes gives the *apparent* power of the circuit, and this must be multiplied by the $\cos\theta$ to give the *actual*

power. This factor $\cos\theta$ is called the *power factor* of the circuit.

When the current and voltage are in-phase, this factor is equal to 1. Resonant or purely resistive circuits are then said to have unity power factor, in which case:

$$W = E \times I,\ W = I^2R,\ W = \frac{E^2}{R}.$$

Applying Ohm's Law to Alternating Current. Ohm's law applies equally to direct or alternating current, *provided* the circuits under consideration are purely resistive, that is, circuits which have neither inductance (coils) nor capacitance (condensers). Problems which involve tube filaments, drop resistors, electric lamps, heaters or similar resistive devices can be solved from Ohm's law, regardless of whether the current is direct or alternating. When a condenser or a coil is made a part of the circuit, a property common to either, called *reactance,* must be taken into consideration.

When the circuit contains inductance only, yet with the same conditions as above, the formula is as follows:

$$E = IX_L, \text{ and } I = \frac{E}{X_L},$$

where E = voltage,
I = current in amperes,
X_L = inductive reactance or $2\pi fL$ (expressed in ohms).

When a circuit has resistance, capacitive reactance, and inductive reactance in *series,* the effective total opposition to the alternating current flow is known as the *impedance* of the circuit. Stated otherwise, impedance of a circuit is the vector sum of the resistance and the difference between the two reactances.

$$Z = \sqrt{r^2 + (X_L - X_c)^2} \text{ or}$$

$$Z = \sqrt{r^2 + \left(2\pi fL - \frac{1}{2\pi fC}\right)}$$

where Z = impedance in ohms,
r = resistance in ohms,
X_L = inductive reactance ($2\pi fL$) in ohms,
X_c = capacitive reactance $\left(\frac{1}{2\pi fC}\right)$ in ohms.

An example will serve to clarify the relationship of resistance and reactance to the total impedance. If a 10-henry choke, a 2-μfd. condenser, and a resistance of 10 ohms (which is represented by the d.c. resistance of the choke) are all connected in *series* across a 60-cycle source of voltage:

for reactance $X_L = 6.28 \times 60 \times 10 = 3{,}750$ ohms (approx.),

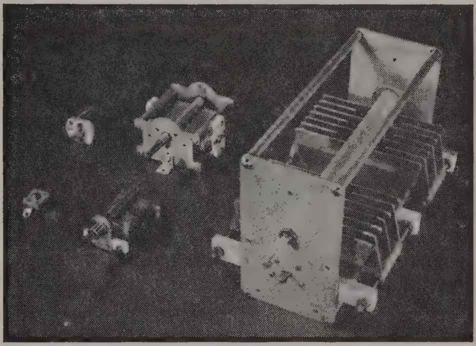
$$X_c = \frac{1{,}000{,}000}{6.28 \times 60 \times 2} = 1{,}300 \text{ ohms (approx.).}$$

r = 10 ohms

Substituting these values in the impedance equation:

$$Z = \sqrt{10^2 + (3750 - 1300)^2} = 2450 \text{ ohms.}$$

This is nearly 250 times the value of the d.c. resistance of 10 ohms. The subject of impedance is more fully covered under *Resonant Circuits.*



Resonant Circuits

The reader is advised to review at this point the subject matter on inductance, capacity, and alternating current, in order that he may gain a complete understanding of the action of resonant circuits. Once the basic conception of the foregoing has been mastered, the more complex circuits in which they appear in combination will present no great problem.

Figure 25 shows an inductance, a capacitance, and a resistance arranged in series, with a variable frequency source, E, of a.c. applied across the combination.

Some resistance is always present in a circuit because it is possessed in some degree by both the inductance and capacitor. If the frequency of the alternator E is varied from nearly zero to some high frequency, there will be one particular frequency at which the inductive reactance and capacitive reactance will be equal. This is known as the *resonant frequency,* and in a series circuit it is the frequency at which the circuit current will be a maximum. Such series resonant circuits are chiefly used when it is desirable to allow a certain frequency to pass through the circuit

Figure 25.
Schematic of a series-resonant circuit containing resistance.

(low impedance to this frequency), while at the same time the circuit is made to offer considerable opposition to currents of other frequencies.

If the values of inductance and capacity both are fixed, there will be only one resonant frequency.

For mechanical reasons, it is more common to change the capacitance rather than the inductance when a circuit is tuned, yet the inductance can be made variable if desired.

In the following table there are five radically different ratios of L to C (inductance to capacitance) each of which satisfies the resonant condition, $X_L = X_C$. When the frequency is constant, L must increase and C must decrease in order to give equal reactance. Figure 26 shows how the two reactances change with frequency; this illustration will greatly aid in clarifying this discussion.

If both the inductance and capacitance are made variable, the circuit may then be changed or *tuned,* so that a number of combinations of inductance and capacitance can resonate at the same frequency. This can be more easily understood when one considers that inductive reactance and capacitive reactance travel in opposite directions as the frequency is changed. For example, if the frequency were to remain constant and the values of inductance and

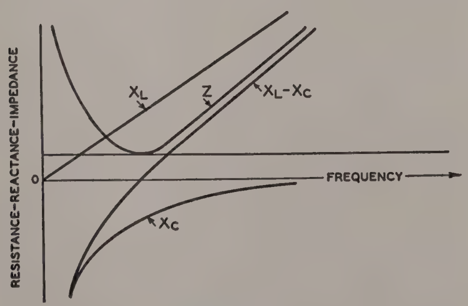


Figure 26.
Variation in reactance and impedance of a series resonant circuit with changing frequency.

capacitance were then changed, the following combinations would have equal *reactance:*

Frequency is constant at 60 cycles.
L is expressed in henrys.
C is expressed in microfarads (.000001 farad.)

L	X_L	*C*	X_C
.265	100	26.5	100
2.65	1,000	2.65	1,000
26.5	10,000	.265	10,000
265.00	100,000	.0265	100,000
2,650.00	1,000,000	.0026	1,000,000

Frequency of Resonance. From the formula for resonance,

$$2\pi fL = \frac{1}{2\pi fC}, \text{ the resonant frequency}$$

can readily be solved. In order to isolate f on one side of the equation, merely multiply both sides by $2\pi f$, thus giving:

$$4\pi^2 f^2 L = \frac{1}{C}.$$

Divided by the quantity $4\pi^2 L$, the result is:

$$f^2 = \frac{1}{4\pi^2 LC}.$$

Then, by taking the square root of both sides:

$$f = \frac{1}{2\pi\sqrt{LC}},$$

where f = frequency in cycles,
L = inductance in henrys,
C = capacity in farads.

It is more convenient to express L and C in smaller units, especially in making radio-frequency calculations; f can also be expressed in megacycles or kilocycles. A very useful group of such formulas is:

$$f^2 = \frac{25{,}330}{LC} \text{ or } L = \frac{25{,}330}{f^2 C} \text{ or } C = \frac{25{,}330}{f^2 L}$$

where f = frequency in megacycles,
L = inductance in microhenrys,
C = capacity in micromicrofarads.

In order to clarify the original formula,

$$f = \frac{1}{2\pi\sqrt{LC}}, \text{ take two values of inductance}$$

and capacitance from the previously given chart and substitute these in the formula. It was stated that the frequency is 60 cycles;

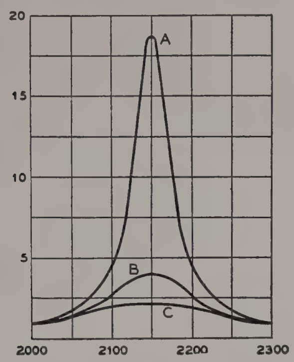


Figure 27.
Resonance curve showing the effect of resistance upon the selectivity of a tuned circuit.

therefore f = 60. Substituting these values to check the frequency:

$$60 = \frac{1}{2\pi\sqrt{LC}};\ 3600 = \frac{1}{4\pi^2 LC};$$

$$L = \frac{1}{3600 \times 4\pi^2 \times .000026}$$

$$L = 0.27$$

The significant point here is that the formula calls for C in *farads,* whereas the capacity was actually in microfarads. Recalling that one microfarad equals .000001 farad, it is, therefore, possible to express 26 microfarads as .000026 farads. This consideration is often overlooked when computing for frequency and capacitive reactance because capacitance is expressed in a totally impractical unit, viz: the *farad.*

Impedance of Series Resonant Circuits. The impedance across the terminals of a series resonant circuit (Figure 25) is:

$$Z = \sqrt{r^2 + (X_L - X_C)^2},$$

where Z = impedance in ohms,
r = resistance in ohms,
X_C = capacitive reactance in ohms,
X_L = inductive reactance in ohms.

From this equation, it can be seen that the impedance is equal to the vector sum of the circuit resistance and the *difference* between the two reactances. Since at the resonant frequency X_L equals X_C, the difference between them (Figure 26) is obviously zero so that at resonance the impedance is simply equal to the resistance of the circuit; therefore, because the resistance of most normal radio-frequency circuits is of a very low order, the impedance is also low.

At frequencies higher and lower than the resonant frequency, the difference between the reactances will be a definite quantity and will add with the resistance to make the impedance higher and higher as the circuit is tuned off the resonant frequency.

Current and Voltage in Series Resonant Circuits. Formulas for calculating currents and voltages in a series resonant circuit are similar to those of Ohm's law.

$$I = \frac{E}{Z} \qquad E = IZ$$

The complete equations:

$$I = \frac{E}{\sqrt{r^2 + (X_L - X_C)^2}}$$

$$E = I\sqrt{r^2 + (X_L - X_C)^2}$$

Inspection of the above formulas will show the following to apply to series resonant circuits: When the impedance is low, the current will be high; conversely, when the impedance is high, the current will be low.

Since it is known that the impedance will be very low at the resonant frequency, it follows that the current will be a maximum at this point. If a graph is plotted of the current against the frequency either side of resonance, the resultant curve becomes what is known as a *resonance curve.* Such a curve is shown in Figure 27.

Several factors will have an effect on the shape of this resonance curve, of which resistance and L-to-C ratio are the important considerations. The curves B and C in Figure 27 show the effect of adding increasing values of resistance to the circuit. It will be seen that the peaks become less and less prominent as the resistance is increased; thus, it can be said that the *selectivity* of the circuit is thereby *decreased.* Selectivity in this case can be defined as the ability of a circuit to discriminate against frequencies adjacent to the resonant frequency.

Voltage Across Coil and Condenser in Series Circuit. Because the a.c. or r.f. voltage across a coil and condenser is proportional to the reactance (for a given current), the actual voltages across the coil and across the condenser may be many times greater than the *terminal* voltage of the circuit. Furthermore, since the individual reactances can be very high, the voltage across the condenser, for example, may be high enough to cause flashover, even though the applied voltage is of a value

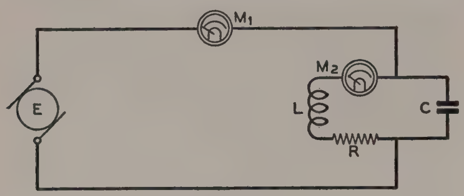


Figure 28.

The parallel resonant (anti-resonant) tank circuit. L and C comprise the reactive elements of the tank and R indicates the initial r.f. resistance of the components. M_1 indicates what is called the "line current" or the current that keeps the tank in a state of oscillation. M_2 indicates the "tank current" or the amount of current circulating through the elements of the tank.

considerably below that at which the condenser is rated.

Circuit Q—Sharpness of Resonance. An extremely important property of a capacitance or an inductance is its factor-of-merit, more generally called its Q. It is this factor, Q, which primarily determines the sharpness of resonance of a tuned circuit. This factor can be expressed as the ratio of the reactance to the resistance, as follows:

$$Q = \frac{2\pi fL}{R},$$

where R = total d.c. and r.f. resistances.

The actual resistance in a wire or inductance can be far greater than the d.c. value when the coil is used in a radio-frequency circuit; this is because the current does not travel through the entire cross-section of the conductor, but has a tendency to travel closer and closer to the surface of the wire as the frequency is increased. This is known as the *skin effect.*

The actual current-carrying portion of the wire is decreased, therefore, and the resistance is increased. This effect becomes even more pronounced in square or rectangular conductors because the principal path of current flow tends to work outwardly toward the four edges of the wire.

Examination of the equation for Q may give rise to the thought that even though the resistance becomes greater with frequency, the inductive reactance does likewise, and that the Q might be a constant. In actual practice, however, this is true only at very low frequencies; the resistance usually increases more rapidly with frequency than does the reactance, with the result that Q normally decreases with increasing frequency.

The Q of a condenser ordinarily is much higher than that of the best coil. Therefore, it usually is the merit of the coil that limits the overall Q of the circuit.

Parallel Resonance. In radio circuits, parallel resonance (more correctly termed *anti-resonance*) is more frequently encountered than series resonance; in fact, it is the basic foundation of receiver and transmitter circuit operation. A circuit is shown in Figure 28.

The "Tank" Circuit. In this circuit, as contrasted with a circuit for series resonance, L (inductance) and C (capacitance) are connected in *parallel,* yet the *combination* can be considered to be in series with the remainder of the circuit. This combination of L and C, in conjunction with R, the resistance which is principally included in L, is sometimes called a *tank* circuit because it effectively functions as a storage tank when incorporated in vacuum tube circuits.

Contrasted with series resonance, there are two kinds of current which must be considered in a parallel resonant circuit: (1) the line current, as read on the indicating meter M_1, (2) the circulating current which flows within the parallel L-C-R portion of the circuit. See Figure 28.

At the resonant frequency, the line current (as read on the meter M_1) will drop to a very low value although the circulating current in the L-C circuit may be quite large. It is this line current that is read by the milliammeter in the plate circuit of an amplifier or oscillator stage of a radio transmitter, and it is because of this that the meter shows a sudden *dip* as the circuit is tuned through its resonant frequency. The current is, therefore, a minimum when a parallel resonant circuit is tuned to resonance, although the *impedance* is a *maximum* at this same point. It is interesting to note that the parallel resonant circuit acts in a distinctly opposite manner to that of a series resonant circuit, in which the current is at a maximum and the impedance is minimum at resonance. It is for this reason that in a parallel resonant circuit the principal consideration is one of impedance rather than current. It is also significant that the *impedance* curve for *parallel* circuits is very nearly identical to that of the *current* curve for *series* resonance. The impedance at resonance is expressed as:

$$Z = \frac{(2\pi fL)^2}{R},$$

where Z = impedance in ohms,
L = inductance in henrys,
f = frequency in cycles,
R = resistance in ohms.

Or, impedance can be expressed as a function of Q as:

$$Z = 2\pi fLQ,$$

showing that the impedance of a circuit is directly proportional to its Q at resonance.

The curves illustrated in Figure 27 can be applied to parallel resonance in addition to the purpose for which they are illustrated.

Reference to the impedance curve will show that the effect of adding resistance to the circuit will result in both a broadening out and a lowering of the peak of the curve. Since the voltage of the circuit is directly proportional to the impedance, and since it is this voltage that is applied to the grid of the vacuum tube in a detector or amplifier circuit, the impedance curve must have a sharp peak in order for the circuit to be *selective*. If the curve is broad-topped in shape, both the desired signal and the interfering signals at close proximity to resonance will give nearly equal voltages on the grid of the tube, and the circuit will then be *non-selective;* i.e., it will tune broadly.

Effect of L/C Ratio in Parallel Circuits. In order that the highest possible voltage can be developed across a parallel resonant circuit, the impedance of this circuit must be very high. The impedance will be greater when the ratio of inductance-to-capacitance is great, that is, when L is large as compared with C. When the resistance of the circuit is very low, X_L will equal X_C at maximum impedance. There are innumerable ratios of L and C that will have *equal* reactance, at a given resonant frequency, exactly as is the case in a series resonant circuit.

In practice, where a certain value of inductance is tuned by a variable capacitance over a fairly wide range in frequency, the L/C ratio will be small at the lowest frequency and large at the high-frequency end. The circuit, therefore, will have unequal gain and selectivity at the two ends of the band of frequencies which is being tuned. Increasing the Q of the circuit (lowering the series resistance) will obviously increase *both* the selectivity and gain.

Circulating Tank Current at Resonance. The Q of a circuit has a definite bearing on the circulating tank current at resonance. This tank current is very nearly the value of the line current multiplied by the circuit Q. For example: an r.f. line current of 0.050 amperes, with a circuit Q of 100, will give a circulating tank current of approximately 5 amperes. From this it can be seen that the inductance and connecting wires in a circuit with a high Q must be of very low resistance, particularly in the case of high power transmitters, if heat losses are to be held to a minimum.

Because the voltage across the tank at resonance is determined by the Q, it is possible to develop very high peak voltages across a high Q tank with but little line current. The high voltage is a result of the heavy circulating current through the tank reactances when the Q is high.

Effect of Coupling on Impedance. If a parallel resonant circuit is coupled to another circuit, such as an antenna output circuit, the impedance of the parallel circuit is decreased as the coupling becomes closer. The effect of closer (tighter) coupling is the same as though an actual resistance were added to the parallel circuit. The resistance thus coupled into the tank circuit can be considered as being *reflected* from the output or load circuit to the driver circuit.

If the load across the parallel resonant tank circuit is purely resistive, just as it might be if a resistor were shunted across the whole tank inductance, the load will not disturb the resonant setting. If, on the other hand, the load is reactive, as it could be with too-long or too-short antenna for the resonant frequency, the setting of the tank tuning condenser will have to be changed in order to restore resonance.

Tank Circuit Flywheel Effect. When the plate circuit of a class B or class C operated tube (defined in the following chapter) is connected to a parallel resonant circuit, the plate current serves to maintain this L/C circuit in a state of oscillation. If an initial impulse is applied across the terminals of a parallel resonant circuit, the condenser will become charged when one set of plates assumes a positive polarity, the other set a negative polarity. The condenser will then discharge through the inductance; the current thus flowing will cut across the turns of the inductance and cause a counter e.m.f. to be set up, charging the condenser in the opposite direction.

In this manner, an alternating current is set up within the L/C circuit and the oscillation would continue indefinitely with the condenser charging, discharging and charging again if it were not for the fact that the circuit possesses some resistance. The effect of this resistance is to dissipate some energy each time the current flows from inductance to condenser and back, so that the amplitude of the oscillation grows weaker and weaker, eventually dying out completely.

The frequency of the initial oscillation is dependent upon the L/C constants of the circuit. If energy is applied in short spurts or pushes at just the right moments, the L/C circuit can be maintained in a constant oscillatory state. The plate current pulses from class B and class C amplifiers supply just the desired kind of kicks.

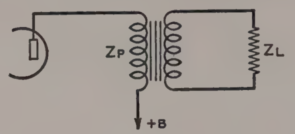


Figure 29.
The reflected impedance Z_P varies directly in proportion to Z_L and in proportion to the square of the turns ratio.

Whereas the class B plate current pulses supply a kick for a longer period, the short pulses from the class C amplifier give a pulse of very high amplitude, thus being even more effective in maintaining oscillation. So it is that the positive half cycle in the tank circuit will be reinforced by a plate current kick, but since the plate current of the tube only flows during a half cycle or less, the *missing* half cycle in the tank circuit must be supplied by the discharge of the *condenser*.

Since the amplitude of this half-cycle will depend upon the charge on the plates of the condenser, and since this in turn will depend upon the capacitance, the value of capacitance in use is very important. Particularly is this true if a distorted wave shape is to be avoided, as would be the case when a transmitter is being modulated. The foregoing applies particularly to single-ended amplifiers. If push-pull were employed, the negative half-cycle would secure an additional kick, thereby greatly lessening the necessity of the use of higher C in the L/C circuit in order to maintain sufficient Q (for a given load resistance).

Impedance Matching: Impedance, Voltage, and Turns Ratio. A fundamental law of electricity is that the maximum transfer of energy results when the impedance of the load is equal to the impedance of the driver. Although this law holds true, it is not necessarily a desirable one for every condition or purpose. In many cases, where a vacuum tube works into a parallel resonant circuit load, it is desirable to have the load impedance considerably higher than the tube plate impedance, so the maximum power will be dissipated by the load rather than in the tube.

Often a vacuum tube circuit requires that the plate impedance of a driver circuit be "matched" to the grid impedance of the tube being driven. When the driven tube operates in such a condition that it draws grid current, such as in all transmitter r.f. amplifier circuits, the grid impedance may well be lower than the plate tank impedance of the driver stage. In this case it becomes necessary to tap down on the driver tank coil in order to select the proper number of turns that will give the desired impedance. If the desired working load impedance of the driver stage is 10,000 ohms, for example, and if the tank coil has 20 turns, the grid impedance of the driven stage being 5000 ohms, it is evident that there will be required a step-down *impedance* ratio of $\frac{10{,}000}{5000}$ or 2-to-1. This impedance value is *not* secured when the driver inductance is tapped at the center. It is of importance to stress the fact that the impedance is decreased *four times* when the number of turns on the tank coil is *halved.* The following equations show this fact:

$$\frac{N_1}{N_2} = \sqrt{\frac{Z_1}{Z_2}} \quad \text{or} \quad \frac{N_1^2}{N_2^2} = \frac{Z_1}{Z_2}$$

where $\frac{N_1}{N_2}$ = turns ratio,

$\frac{Z_1}{Z_2}$ = impedance ratio.

In the foregoing example, a step-down *impedance* ratio of 2-to-1 would require a *turns* step-down ratio of the square root of the impedance, or 1.41. Therefore, if the inductance has 20 turns, a tap would be taken on the sixth turn down from the hot end, or 14 turns up from the cold end. This is arrived at by taking the resultant for the turns ratio, i.e., 1.41, and then dividing it into the total number of turns, as follows:

$$\frac{20}{1.41} = (14 \text{ approx.}).$$

Either an impedance step-up or step-down ratio can be secured from a parallel resonant circuit. One type of antenna impedance matching device utilizes this principle. Here, however, two condensers are effectively in series across the inductance; one has quite a high capacitance (500 $\mu\mu$fd.), the other is a conventional size condenser used principally to restore resonance. The theory of the device is simply that the impedance is proportional to the reactances of the condensers, and, by changing the ratio of the two, the antenna is effectively connected into the tank circuit at impedance points which reach higher or lower values as the ratio of the condensers is changed.

As the impedance step-down ratio becomes larger, the voltage step-down becomes correspondingly great. Such a condition takes place when a resonant circuit is tapped down for reasons of impedance matching; the *voltage*

will be stepped down in *direct proportion* to the *turn* step-down ratio. The reverse holds true for step-up ratios. As the step-up ratio is increased, the voltage is increased.

The foregoing discussion ignores the effect of *leakage reactance,* or, in other words, the fact that there is not perfect coupling between different parts of the coil. However, in the case of a well designed coil, the effects are not large, and may be ignored for this discussion.

Transformers

When two coils are placed in such inductive relation to each other that the lines of force from one cut across the turns of the other and induce a voltage in so doing, the combination can be called a *transformer.* The name is derived from the fact that energy is transformed from one voltage into another. The inductance in which the original flux is produced is called the *primary;* the inductance which *receives* the induced voltage is called the *secondary.* In a radio receiver power transformer, for example, the coil through which the 110-volt a.c. passes is the *primary,* and the coil from which a higher or lower voltage than the a.c. line potential is obtained is the *secondary.*

Transformers can have either air or magnetic cores, depending upon whether they are to be operated at radio or audio frequencies. The reader should thoroughly impress upon his mind the fact that current can be transferred from one circuit to another *only* if the primary current is changing or alternating. From this it can be seen that a power transformer cannot possibly function as such when the primary is supplied with non-pulsating d.c.

A power transformer usually has a magnetic core which consists of laminations of iron, built up into a square or rectangular form, with a center opening or window. The secondary windings may be several in number, each perhaps delivering a different voltage. The secondary voltages will be proportional to the number of turns and to the primary voltage.

If a primary winding has an a.c. potential of 110 volts applied to 220 turns of wire on the primary, it is evident that this winding will have 2 turns per volt. A secondary winding of 10 turns, wound on an adjacent leg of the transformer core, would have a potential of 5 volts. If the secondary winding has 500 turns, the potential would be 250 volts, etc. Thus, a transformer can be designed to have either a step-up or step-down ratio, or both simultaneously. The same applies to air core transformers for radio-frequency circuits.

Transformer Action. Transformers are used in alternating current circuits to transfer power at one voltage and impedance to another circuit at another voltage and impedance. There are three main classifications of transformers: those made for use in power-frequency circuits (25, 50, and 60 cycles), those made for use at radio frequencies, and those made for audio-frequency applications. Power transformers will be discussed in the section devoted to *Power Supplies,* and r.f. transformers are analyzed later on in this chapter; a few of the pertinent facts concerning audio transformers will be covered in the following paragraphs.

Impedance Matching in Audio Circuits. In most audio applications it will be the function of the audio transformer to match the impedance of the plate circuit of a vacuum-tube amplifier to a load circuit of a different impedance. The information given under the paragraph headed *Impedance Matching* is very easily applied to this type of calculation.

In all audio-frequency circuit applications, it is only necessary to refer to the *tube tables* in this book in order to find the recommended load impedance for a given tube and a given set of operating conditions. For example, the table shows that a type 42 pentode tube requires a load impedance of 7000 ohms. Audio transformers are always rated for both their primary and secondary impedance, which means that the primary impedance will be of the rated value *only* when the secondary is terminated in its rated impedance.

If a 7000-ohm plate load is to work into a 7-ohm loudspeaker voice coil, the impedance ratio of the transformer would be $\frac{7000}{7}$ = 1000-to-1. Hence, the turns-ratio will be the square root of 1000, or 31.6. This does not mean that the primary will have only 31.6 turns of wire and only 1 turn on the secondary. The primary must have a certain *inductance* in order to offer a high impedance to the lower audio frequencies. Consequently, it must have a large number of turns of wire in the primary winding. The *ratio* of total primary turns to total secondary turns must remain constant, regardless of the number of turns in the primary if the correct primary impedance is to be maintained.

To summarize, a certain transformer will have a certain impedance ratio (determined by the square of the turns ratio) which will remain constant. If the transformer is terminated with an impedance or resistance *lower* than the original rated value, the reflected impedance on the primary will also be lower

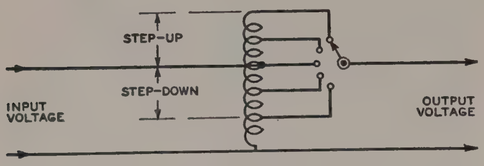


Figure 30.
Schematic diagram of an auto-transformer showing the method of connecting it to the line and to the load.

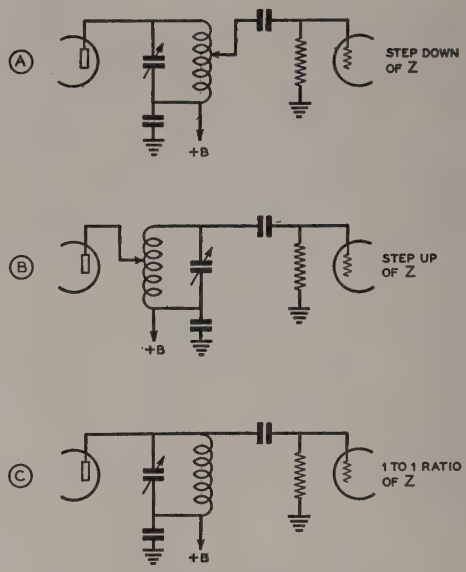


Figure 31.
Impedance step-up and step-down may be obtained by utilizing the plate tank circuit of a vacuum tube as an auto-transformer.

than the rated value. If the transformer is terminated in an impedance *higher* than rated, the reflected primary impedance will be higher.

For push-pull amplifiers the recommended primary impedance is stated as some certain value, *plate to plate;* this refers to the impedance of the total winding without consideration of the center tap. The reflected impedance across the total primary will follow the same rules as previously given for single-ended stages.

The voltage relationship in primary and secondary is the same as the turns ratio. For a step-down turns ratio of 10-to-1, the corresponding *voltage* step-down would be 10-to-1 though the *impedance* ratio would be 100-to-1. This information is useful when it is desired to convert the turns ratios given on certain types of driver transformers into impedance ratios.

The same type of reasoning and subsequent calculation would be used in determining the turns ratio for a modulation transformer to couple a certain pair of class B modulators to a class C final amplifier. The recommended plate-to-plate load impedance for the modulator tubes can be obtained from the tube tables given later on. The final amplifier load resistance is then determined by dividing its plate voltage by the plate current at which it is to operate. The turns ratio of the modulation transformer is then equal to the square root of the ratio between the modulator load impedance and the amplifier load resistance; the transformer may be either step-up or step-down, as the case may be.

The Auto Transformer. The type of transformer in Figure 30, when wound with heavy wire over an iron core, is a common device in primary power circuits for the purpose of increasing or decreasing the line voltage. In effect, it is merely a continuous winding with taps taken at various points along the winding, the input voltage being applied to the bottom and also to one tap on the winding. If the output is taken from this same tap, the voltage ratio will be 1-to-1; i.e., the input voltage will be the same as the output voltage. On the other hand, if the output tap is moved down toward the common terminal, there will be a step-down in the turns ratio with a consequent step-down in voltage.

The opposite holds true if the output terminal is moved upward from the middle input terminal; there will be a voltage step-up in this case. The initial setting of the middle input tap is chosen so that the number of turns will have sufficient reactance to keep the no-load primary current at a reasonably low value.

In the same manner as voltage is stepped up and down by changing the number of turns in a winding, so can impedance be stepped up or down. Figure 31A shows an application of this principle as applied to a vacuum tube circuit which couples one circuit to another.

Assuming that the grid impedance may be of a lower value than the plate tank impedance (desired load impedance) of the preceding stage, a step-down ratio will be necessary in order to give maximum transfer of energy. In B of Figure 31, the grid impedance is very high as compared with the tank impedance of the driver stage, and thus there is required a step-up ratio to the grid. The driver plate is

tapped down on its plate tank coil in order to make this impedance step-up possible. A driver tube with very low plate impedance must be used if a good order of plate efficiency is to be realized.

In C of Figure 31, the grid impedance very closely approximates the desired plate load impedance, and this connection is used when no transformation is required.

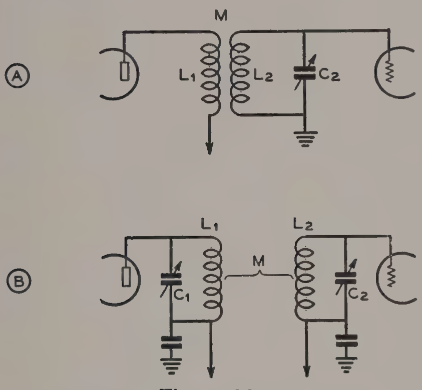


Figure 32.
Two commonly used types of inductive coupling between radio-frequency circuits.

Inductive Coupling — The Radio-Frequency Transformer. Inductive coupling is often used when two circuits are to be coupled. This method of coupling is shown in Figures 32A and 32B.

The two inductances are placed in such inductive relation to each other that the lines of force from the primary coil cut across the turns of the secondary coil, thereby inducing a voltage in the secondary. As in the case of capacitive coupling, impedance transformation here again becomes of importance. If two parallel tuned circuits are coupled very closely together, the circuits can in reality be overcoupled. This is illustrated by the curve in Figure 33.

The dotted line, curve A, is the original curve or that of the primary coil *alone.* Curve B shows what takes place when two circuits are overcoupled; the resonance curve will have a definite dip on the peak, or a double hump. This principle of overcoupling is advantageously utilized in bandpass circuits where, as shown in C, the coupling is adjusted to such a value as to reduce the peak of the curve to a virtual flat top, with no dip in the center as in B.

Some undesirable capacitive coupling may

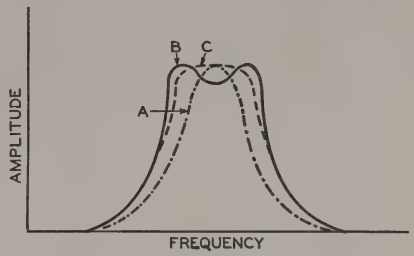


Figure 33.
Effect of coupling between circuits upon the resonance curve. Curve A indicates the curve when the circuits are under-coupled, B is the curve resulting from over-coupling, and C is the curve resulting from an intermediate value of coupling.

result when circuits are closely or tightly coupled; if this capacitive coupling is appreciable, the tuning of the circuits will be affected. The amount of capacitive coupling can be reduced by so arranging the physical shape of the inductances as to enable only a minimum surface of one to be presented to the other.

Another method of accomplishing the same purpose is by electrical means. A curtain of closely-spaced parallel wires or bars, connected together only at one end, and with this end connected to ground, will allow electromagnetic coupling but not electrostatic coupling. Such a device is called a *Faraday screen.*

Link Coupling. Still another method of decreasing capacitive coupling is by means of a *coupling link* circuit between two parallel resonant circuits. The capacity of the coupling link, with its 1 or 2 turns, is so small as to be negligible. Also, one side of the link is often grounded to reduce further any capacitive coupling that may exist.

Link coupling is widely used in transmitter circuits because it adapts itself so universally and eliminates the need of a radio-frequency choke, thereby reducing a source of loss. Link coupling is very simple; it is diagrammed in A and B of Figure 34.

In A of Figure 34, there is an impedance step-down from the primary coil to the link circuit. This means that the line which connects the two links or loops will have a low impedance and therefore can be carried over a considerable distance without introduction of appreciable loss. A similar link or loop is at the output end of the line; this loop is coupled to the grid tank of the driven stage.

Still another link coupling method is shown in B of Figure 34. It is similar to that of A, with the exception that the primary line is

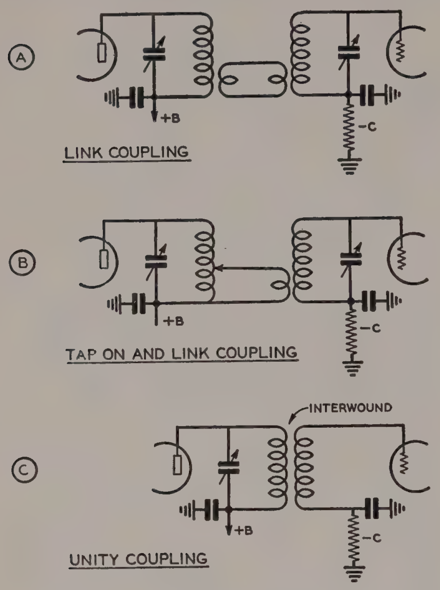


Figure 34.
Two types of link (inductive) coupling and (C) unity coupling.

tapped on the coil, rather than being terminated in a link or loop.

Unity Coupling. Another commonly used type of coupling is that known as *unity coupling,* by reason of the fact that the turns ratio between primary and secondary is 1-to-1. This method of coupling is illustrated in C of Figure 34. Only one of the windings is tuned, although the interwinding of the two coils gives an effect in the untuned winding as though it were actually tuned with a condenser.

Unity coupling is used in some types of ultra-high-frequency circuits, although the mechanical considerations are somewhat difficult. The secondary, when serving as a grid coil, is placed inside of a copper tubing coil; the latter serves as the primary or plate coil.

Conduction of an Electric Current

So far this chapter has dealt only with the conduction of current by a stream of electrons through a conductor or by electrostatic coupling through a capacitor. While this is the most common method of transmission, there are other types of conduction which are equally important in their respective branches of the field. An electric current may also be transmitted by the motion of minute particles of matter, by the motion of charged atoms called *ions,* and by a stream of electrons in a vacuum.

The carrying of current by charged particles, such as bits of dust, is only of academic interest in radio. However, there is a commercial process (called the Cottrell process) which uses this type of conduction in industrial dust precipitation. A highly charged wire inside a grounded metal chamber is placed so that the dust-laden flue gases from certain industrial processes (usually metallurgic refining) must pass through the chamber. The dust particles are first attracted to the wire; there they attain a high electric charge which causes them to be attracted to the sides of the chamber where they are precipitated and subsequently collected. A small electric current between the center electrode and the chamber is the result of the carrying of the charges by the dust particles.

Conduction by Ions. When a high enough voltage is placed between two terminals in air or any other gas, that gas will break down suddenly, the resistance between the two points will drop from an extremely high value to a very low value, and a comparatively large electric current will flow to the accompaniment of an amount of visible light either as a flash, an arc, a spark, or a colored discharge such as is found in the "neon" sign. This type of conduction is due to gas ions which are generated when the electric stress between the two points becomes so great that electrons are torn from the molecules of the gas with the production of a quantity of positively charged gas ions and negative electrons. The breakdown voltage for a particular gas is dependent upon the pressure, the spacing of the electrodes, and the type of electrodes.

Lightning, tank condenser flashovers, and ignition sparks in an automobile are such discharges that occur at atmospheric pressure or above. However, the pressure of the gas is usually reduced to facilitate the ease of breakdown of the gas, as in the "neon" sign, mercury-vapor lamp, or voltage regulator tubes such as the VR-150-30. If a heated filament is used as one electrode in the discharge chamber, the breakdown voltage is further reduced to a value called the *ionization potential* of the gas. This principle is used in the 866, the 83, and other mercury-vapor rectifiers. Through the use of the heated cathode, the break-down potential is reduced from about 10,000 volts to approximately 15 volts and the conduction of electric current is made unidirectional, enabling the discharge chamber to be used as a rectifier. The applications of the principle of ionic conduction in vacuum tubes (along with discussion of

electronic conduction) will be covered in more detail in the chapter devoted to *Vacuum Tube Theory*.

The emission of colored light which accompanies an electric discharge through a gas is due to the re-combination of the ionized gas molecules and the free electrons to form neutral gas molecules. There is a definite color spectrum which is characteristic of every gas—and for that matter for every element when it is in the gaseous state. For neon this color is orange-red, for mercury it is blue-violet, for sodium, almost pure yellow—and so on through the list of the elements. This principle is used in the spectroscopic identification of elements by their characteristic lines in the spectrum (called Fraunhöfer lines).

Electrolytic Conduction. Nearly all inorganic chemical compounds (and a few organic ones of certain molecular structure) when dissolved in water undergo a chemical-electrical change known as *electrolytic disassociation* which results in the production of ions similar in certain properties to those formed as a result of the electric breakdown of a gas. For example, when sodium chloride or table salt is dissolved in water a certain percentage of it ionizes or breaks down into positively charged sodium ions, or sodium atoms with a deficiency of one electron, and negatively charged chloride ions, or chlorine atoms with one excess electron. Similarly, sodium hydroxide disassociates into positive sodium ions and negative hydroxyl ions—sulfuric acid into positive hydrogen ions and negative sulfate ions.

This solution of an ionized compound and water renders the aqueous solution a conductor of electricity. (Water in the pure form is a good insulator.) The conductivity of the solution is proportional to the mobility of the ions and to the quantity of them available in the solution. Maximum conductivity is had not when there is a maximum of the compound in solution but rather when there is a maximum of ions in solution; this condition is ordinarily obtained when neither concentrated nor dilute but about midway between. Maximum conductivity in a sulfuric acid solution as used in storage batteries is obtained when there is about 30 per cent by weight of the acid in solution in the water. It is for this reason that acid of about 30 per cent concentration is used as an electrolyte in storage batteries.

Conduction of electricity through an *electrolyte,* as a conducting solution is called, is made possible by the mobility of the charged ions in solution. When a positively and a negatively charged wire are placed in an electrolyte the negative ions are attracted to the positive wire and the positive ions are attracted to the negative wire. As the ions reach the wire carrying the charge opposite to their own, their excess or their deficiency of electrons is neutralized by the respective deficiency or excess of electrons on the wire, and the ion changes from the ionic to the atomic or molecular state. If the ion happened to be that of a metal such as copper, copper will be *plated* upon the negative electrode that had been placed into the solution; if the negative ion was that of chlorine (the chloride ion), then chlorine in the gaseous form will appear at the positive electrode. The conduction of an electric current through an electrolyte always results in a chemical change in the electrolyte. This fact is employed commercially in electroplating and electrolytic refining processes.

The Primary Cell. If two dissimilar metals are placed in an electrolyte a potential difference will appear between the two materials. This postulate is employed commercially in the primary cell, or "dry cell" as it is somewhat incorrectly called.

The operation of the primary cell depends upon the differences in the two electrochemical constants for the materials used as the electrodes. With the zinc and carbon used in the dry cell (with a paste containing ammonium chloride as the electrolyte) the potential is 1.53 volts. With other electrolytes and electrodes the potential output of the cell varies from 0.7 to 2.5 volts.

When current is taken from a primary cell, the negative electrode (usually the zinc container) dissolves in the electrolyte with the production of hydrogen gas. If only the positive and the negative electrodes and the electrolyte were contained in the cell, this hydrogen gas would collect as a film on the surface of the negative electrode. When this film does form, the internal resistance of the cell increases due to the insulating properties of the film of gas. A cell is said to have become "polarized" when this has taken place. To reduce this effect, an oxidizing agent called a "depolarizer" (manganese dioxide, in the case of the dry cell) is incorporated into the electrolyte. If current is taken from the cell at a reasonable rate the depolarizer oxidizes the hydrogen into water as fast as it is formed. This formation of water as a result of the normal operation of the cell is one of the reasons that a dry cell "sweats" when it is approaching the end of its useful life.

Dry cells and batteries of them are very commonly employed in portable radio equipment as both filament and plate supply, and frequently as plate supply at locations where there is no source of alternating current.

Through recent improvements in cell manufacture and in the design of batteries of these cells it is possible to make very lightweight sources of a quite reasonable amount of power. 45-volt B batteries are available ranging in weight from 16 pounds down to about 2 ounces. The large sizes will stand current drain up to about 75 ma. for a few hundred hours while the smallest sizes will last only a few hours with a drain of 1 or 2 milliamperes. Medium sizes capable of producing 8 to 10 ma. for 100 hours or so are commonly used in radio-controlled model aircraft and in the new portable broadcast receivers.

The Secondary Cell—Storage Batteries. The primary cell, as described in the preceding paragraphs, produces its voltage as a result of chemical action of the electrolyte on one of the elements. When the material comprising the active element is used up, the cell is no longer useful and must be discarded. The secondary cell, on the other hand, is capable of being recharged to its original energy content when it has been depleted.

There are two common types of secondary cells: the *Edison cell,* which uses iron as the negative pole and nickel oxide as the positive in a 20 per cent solution of potassium hydroxide as the electrolyte; and the *lead cell,* which uses lead as the negative pole and lead dioxide as the positive pole in an electrolyte of 30 per cent sulfuric acid.

The Edison cell battery has longer life, and generally will stand more abuse than a lead-acid type battery. However, the lower cost of the latter type battery makes it much more widely used. The common automobile battery is a lead-acid type battery. The lead-acid type battery has a much lower internal resistance, which makes it much more suitable where heavy current must be delivered for a short time.

When a storage battery is being discharged, chemical energy is being converted into electrical energy. When the battery is being charged, by causing a reverse current to flow between electrodes, electrical energy is being converted to chemical energy. Actually, a battery cannot "store" electricity; only a condenser can do that.

CHAPTER THREE

Vacuum-Tube Theory

THE science of radio is based upon one of the most versatile developments of the twentieth century—the electron tube, or as it is more commonly named, the vacuum tube. It is the utilization of the unique characteristics of the vacuum tube in various circuit arrangements which makes possible modern radio communication; for that matter, long distance wire communication also owes its efficiency to the versatility of the vacuum tube.

This chapter is divided into two main sections. The first is devoted to the basic theory of the vacuum tube and to a discussion of the various types of tubes which have been developed up to the present time. The second part discusses the application of the vacuum tube to the various circuit arrangements which have been developed to utilize its characteristics.

Brief History of the Vacuum Tube. Thomas Edison is credited with the discovery that an additional wire or plate placed inside a lighted incandescent lamp would acquire a negative charge of electricity. J. A. Fleming undertook the study of the *Edison Effect* in 1895, and as a result of his findings in 1904 he patented the two-electrode tube or *diode* which became known as the Fleming valve. Then, in 1906, Lee de Forest discovered that a third element could be placed between the cathode and plate to control the flow of electrons from one to the other. This third element was called a *grid* from its physical resemblance to the grid or grate of a stove. The insertion of the grid into the space between the cathode and plate in the diode resulted in the most versatile of vacuum tubes, the *triode*.

In recent times other elements or grids have been added to the original triode to augment the electron flow in a particular manner, or to give a particular characteristic to the vacuum tube. These later types have been called *multi-element* tubes. The names for these multi-element tubes are obtained by adding the Greek prefix for the number of elements to the root *-ode:* diode, triode, tetrode, pentode, hexode, and heptode respectively for tubes having two, three, four, five, six, and seven elements.

MECHANICS OF THE VACUUM TUBE

The original Edison discovery was that a heated filament would give off electrons to a cold plate in the same evacuated chamber. It was later discovered that if the plate were charged positively with respect to the filament, a much larger proportion of the emitted electrons would be attracted to the plate. But, if the plate were charged negatively with respect to the filament, the electron flow would stop. This valve action meant that the vacuum tube could be used as a rectifier since it would pass current only in one direction. It is this rectifying action of the diode which is used almost universally for the production of direct current from alternating current as supplied by the a.c. mains.

Then, the discovery that additional elements could be placed between the cathode and plate to control the electron flow in any desired manner resulted in the simultaneous development of the vacuum tube and improvement of the radio art to make use of the greater capabilities of the improved tubes. In recent years, however, the improvement of the art has commonly been first, with improved types of vacuum tubes being developed as the need for them arose.

Thermionic Emission. The free electrons in any metal are continually in motion at all temperatures. But at ordinary atmospheric temperatures, these electrons do not have sufficient energy to penetrate the surface of the material. It is necessary that some form of external energy be supplied to the surface for emission to take place. When this energy supply is in the form of heat, the result is called thermionic emission; when the energy is in the form of light it is called photo-emission. The phenomena of photo-emission is applied in the photo-electric tube, while thermionic emission

supplies the electrons for the operation of the vacuum tube.

In order that thermionic emission may take place, it is necessary that the cathode or filament of the vacuum tube be heated to the point where the free electrons in the emitter have sufficient velocity to penetrate the surface. The degree of temperature to which the emitter must be heated varies greatly with the type of emitter. Since there are several types of emitters commonly found in present day transmitting and receiving tubes, these will be described separately.

Types of Emitters

Emitters as used in present-day vacuum tubes may be classed into two groups: the directly heated or filament type, and the indirectly heated or heater-cathode type.

Directly heated emitters may be further subdivided into three important groups, all of which are important and commonly used in modern tubes. These classifications are: the pure tungsten filament, the thoriated-tungsten filament, and the oxide-coated filament.

The Pure Tungsten Filament. Pure tungsten wire was used as the filament in nearly all the earlier transmitting and receiving tubes. However, the thermionic efficiency of tungsten wire as an emitter (the number of milliamperes emission per watt of filament heating power) is quite low, the filaments become fragile after use, their life is rather short, and they are susceptible to burnout at any time. Pure tungsten filaments must be run at bright white heat (about 2500° Kelvin). For these reasons, tungsten filaments have been replaced in all applications where another type of filament could be used. They are, however, still universally employed in most water-cooled tubes and in certain large, high-power air-cooled triodes where another filament type would be unsuitable. Tungsten filaments are the most satisfactory for high-power, high-voltage tubes where the emitter is subjected to positive ion bombardment due to the residual gas content of the tubes. Tungsten is not adversely affected by such bombardment.

The Thoriated-Tungsten Filament. In the course of experiments made upon tungsten emitters, it was found that filaments made from tungsten having a small amount of thoria (thorium oxide) as an impurity had much greater emission than those made from the pure metal. Subsequent improvements have resulted in the highly efficient carburized thoriated-tungsten filament as used in virtually all medium-power transmitting tubes today.

Thoriated-tungsten emitters consist of a tungsten wire containing about 1 per cent thoria. The new filament is first carburized by heating it to a high temperature in an atmosphere containing a hydrocarbon at reduced pressure. Then the envelope is highly evacuated and the filament is flashed for a minute or two at about 2600° K before being burned at 2200° K for a longer period of time. The flashing causes some of the thoria to be reduced by the carbon to metallic thorium. The activating at a lower temperature allows the thorium to diffuse to the surface of the wire to form a layer of the metal one molecule thick. It is this single-molecule layer of thorium which reduces the work function of the tungsten filament to such a value that the electrons will be emitted from a thoriated filament thousands of times more rapidly than from a pure tungsten filament *operated at the same temperature.*

The carburization of the tungsten surface seems to form a layer of tungsten carbide which holds the thorium layer much more firmly than the plain tungsten surface. This allows the filament to be operated at a higher temperature, with consequent greater emission, for the same amount of thorium evaporation. Thorium evaporation from the surface is a natural consequence of the operation of the thoriated-tungsten filament. The carburized layer on the tungsten wire plays another role in acting as a reducing agent to produce new thorium from the thoria to replace that lost by evaporation. This new thorium continually diffuses to the surface during the normal operation of the filament.

One thing to remember about any type of filament, particularly the thoriated type, is that the emitter deteriorates practically as fast when "standing by" (no plate current) as it does with any normal amount of emission load. However, a thoriated filament may be either temporarily or permanently damaged by a heavy overload which may strip the surface emitting layer of thorium from the filament.

Reactivating Thoriated-Tungsten Filaments. Thoriated-tungsten filaments (and *only* thoriated-tungsten filaments) which have gone "flat" as a result of insufficient filament voltage, a severe temporary overload, a less severe extended overload, or even normal operation may quite frequently be reactivated to their original characteristics by a process similar to that of the original activation. However, only filaments which have been made by a reputable manufacturer and which have not approached too close to the end of their useful life may be successfully reactivated. The filament found in certain makes of tubes may often be reactivated three or four times before

the filament will cease to operate as a thoriated emitter.

The actual process of reactivation is simple enough and only requires a filament transformer with taps allowing voltage up to about 25 volts or so. The tube which has gone flat is placed in a socket to which only the two filament wires have been connected. The filament is then "flashed" for about 20 to 40 seconds at from $1\frac{1}{2}$ to 2 times normal rated voltage. The filament will become extremely bright during this time and, if there is still some thoria left in the tungsten and if the tube didn't originally fail as a result of an air leak, some of this thoria will be reduced to metallic thorium. The filament is then burned at 15 to 25 per cent overvoltage for from 30 minutes to 3 to 4 hours to bring this new thorium to the surface.

The tube should then be tested to see if it shows signs of renewed life. If it does, but is still weak, the burning process should be continued at about 10 to 15 per cent overvoltage for a few more hours. This should bring it back almost to normal. If the tube checks still very low after the first attempt at reactivation, the complete process can be repeated as a last effort.

Thoriated-tungsten filaments are operated at about 1900° K or at a bright yellow heat. A burnout at normal filament voltage is almost an unheard of occurrence. The ratings placed upon tubes by the manufacturers are figured for a life expectancy of 1000 hours. Certain types of tubes may give much longer life than this but the average transmitting tube will give from 1000 to 5000 hours of useful life.

The Oxide-Coated Filament. The most *efficient* of all modern filaments is the oxide-coated type which consists of a mixture of barium and strontium oxides coated upon a wire or strip usually consisting of a nickel alloy. This type of filament operates at a dull-red to orange-red temperature (1050° to 1170° K) at which temperature it will emit large quantities of electrons. The oxide-coated filament is somewhat more efficient than the thoriated-tungsten type in small sizes and it is considerably less expensive to manufacture. For this reason all receiving tubes and quite a number of the low-powered transmitting tubes use the oxide-coated filament. Another advantage of the oxide-coated emitter is its extremely long life—the average tube can be expected to run from 3000 to 5000 hours, and when loaded very lightly, tubes of this type have been known to give 50,000 hours of life before their characteristics changed to any great extent.

The oxide-coated filament does have the disadvantage, however, that it is unsuitable for use in tubes which must withstand more than about 600 volts of plate potential. Some years back, transmitting tubes for operation up to 2000 volts were made with oxide-coated filaments but they have been discontinued. Much more satisfactory operation is obtainable at medium plate potentials with thoriated filaments.

Oxide filaments are unsatisfactory for use at high plate voltages because: (1) their activity is seriously impaired by the high temperature necessary to de-gas the high-voltage tubes and, (2) the positive ion bombardment which takes place even in the best evacuated high-voltage tube causes destruction of the oxide layer on the surface of the filament.

Oxide-coated filaments operate by virtue of a mono-molecular layer of alkaline-earth metal (barium and strontium) which forms on the surface of the oxide coating. Such filaments do not require reactivation since there is always sufficient reduction of the oxides and diffusion of the metals to the surface of the filament to more than meet the emission needs of the cathode.

Indirectly Heated Filaments—The Heater Cathode

The heater type cathode was developed as a result of the requirement for a type of emitter which could be operated from alternating current and yet would not introduce a.c. ripple modulation even when used in low-level stages. It consists essentially of a small nickel-alloy cylinder with a coating of strontium and barium oxides on its surface similar to that used on the oxide-coated filament. Inside the cylinder is an insulated heater element consisting usually of a double spiral of tungsten wire. The heater may operate on any voltage from 2 to 117 volts, although 6.3 is by far the most common value. The heater is operated at quite a high temperature so that the cathode itself may be brought to operating temperature in a matter of 15 to 30 seconds. Heat coupling between the heater and the cathode is mainly by radiation, although there is some thermal conduction through the insulating coating on the heater wire, as this coating is also in contact with the cathode thimble.

Indirectly heated cathodes are employed in all a.c. operated tubes which are designed to operate at a low level either for r.f. or a.f. use. However, some receiver power tubes use heater cathodes (6L6, 6V6, 6F6, and 6B4G) as do some of the low-power transmitter tubes (802, 807, T21, and RK39). Heater cathodes are employed exclusively when a number of tubes are to be operated in series as in an a.c.-

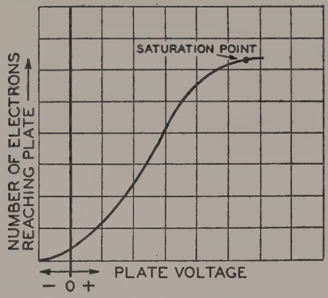


Figure 1.
CURVE SHOWING NUMBER OF ELECTRONS REACHING THE PLATE OF A DIODE PLOTTED AS A FUNCTION OF THE PLATE VOLTAGE.

It will be noticed that there is a small flow of plate current even with zero voltage. This initial flow can be stopped by a small negative plate potential. As the plate voltage is increased in a positive direction, the plate current increases approximately as the 3/2 power of the plate voltage until the saturation point is reached. At this point all the electrons being emitted from the cathode are being attracted to the anode.

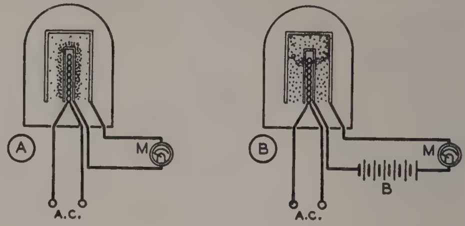


Figure 2.
ILLUSTRATING THE SPACE CHARGE EFFECT IN A DIODE.

(A) shows the space charge existing in the vicinity of the cathode with zero or a small amount of plate voltage. A few high-velocity electrons will reach the plate to give a small plate current even with no plate voltage. (B) shows how the space charge is neutralized and all the electrons emitted by the cathode are attracted to the plate with a battery sufficient to cause saturation plate current.

d.c. receiver. A heater cathode is often called a uni-potential cathode because there is no voltage drop along its length as there is in the filament-type cathode.

Types of Vacuum Tubes

If a cathode capable of being heated either indirectly or directly is placed in an evacuated envelope along with a plate, such a two-element vacuum tube is called a diode. The diode is the simplest of all vacuum tubes and is the fundamental type from which all the others are derived; hence, the diode and its characteristics will be discussed first.

Characteristics of the Diode. When the cathode within a diode is heated, it will be found that a few of the electrons leaving the cathode will leave with sufficient velocity to reach the plate. If the plate is electrically connected back to the cathode, the electrons which have had sufficient velocity to arrive at the plate will flow back to the cathode through the external circuit. This small amount of initial plate current is an effect found in all two-element vacuum tubes.

If a battery or other source of d.c. voltage is placed in the external circuit between the plate and cathode so that the battery voltage places a positive potential on the plate, the flow of current from the cathode to plate will be increased. This is due to the strong attraction offered by the positively charged plate for any negatively charged particles. If the positive potential on the plate is increased, the flow of electrons between the cathode and plate will also increase up to the point of *saturation*. Saturation current flows when all of the electrons leaving the cathode are attracted to the plate, and no increase in plate voltage can increase the number of electrons being attracted.

The Space Charge Effect. As a cathode is heated so that it begins to emit, those electrons which have been discharged into the surrounding space form in the immediate vicinity of the cathode a negative charge which acts to repel those electrons which normally would be emitted were the charge not present. This cloud of electrons around the cathode is called the *space charge*. The electrons comprising the charge are continuously changing, since those electrons making up the original charge fall back into the cathode and are replaced by others emitted by it.

The effect of the space charge is to make the current through the tube variable with respect to the plate-to-cathode drop across it. As the plate voltage is increased, the positive charge of the plate tends to neutralize the negative space charge in the vicinity of the cathode. This neutralizing action upon the space charge by the increased plate voltage allows a greater number of electrons to be emitted from the cathode which, obviously, causes a greater plate current to flow. When the point is reached at which the space charge around the cathode is neutralized completely,

all the electrons that the cathode is capable of emitting are being attracted to the plate and the tube is said to have reached *saturation* plate current as mentioned above.

Insertion of a Grid—The Triode. If an element consisting of a mesh or spiral of wire is inserted concentric with the plate and between the plate and the cathode, such an element will be able to control by electrostatic action the cathode-to-plate current of the tube. The new element is called a grid, and a vacuum tube containing a cathode, grid, and plate is commonly called a triode.

If this new element through which the electrons must pass in their course from cathode to plate is made negative with respect to the cathode, the negative charge on this grid will effectively repel the negatively charged electrons (like charges repel; unlike charges attract) back into the space charge surrounding the cathode. Hence, the number of electrons which are able to pass through the grid mesh and reach the plate will be reduced, and the plate current will be reduced accordingly. As a matter of fact, if the charge on the grid is made sufficiently negative, all the electrons leaving the cathode will be repelled back to it and the plate current will be reduced to zero. Any d.c. voltage placed upon a grid is called a *bias* (especially so when speaking of a control grid). Hence, the smallest negative voltage which will cause cutoff of plate current with a particular plate voltage is called the value of *cutoff bias*.

Figure 3 illustrates an analogy of the method in which the number of electrons flowing to the plate is controlled by the grid bias. Figure 4 graphically shows essentially the same information as shown in Figure 3; i.e., the manner in which the plate current of a typical triode will vary with different values of grid bias. Figure 4 also shows graphically the cutoff point, the approximately linear relation between grid bias and plate current over the operating range of the tube, and the point of plate current saturation. However, the point of plate current saturation comes at a different position with a triode as compared to a diode. Plate current non-linearity or saturation may begin either at the point where the full emission capabilities of the filament have been reached, or at the point where the positive grid voltage begins to approach the positive plate voltage.

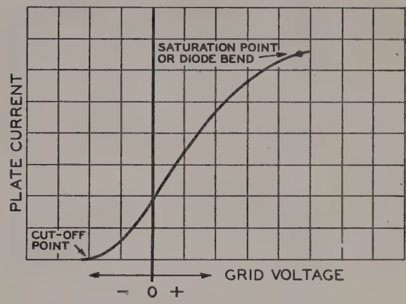


Figure 4.

PLATE CURRENT PLOTTED AGAINST GRID VOLTAGE, WITH CONSTANT PLATE VOLTAGE.

For values of grid bias between those which give plate current cutoff and plate current saturation, the value of plate current varies more or less linearly with respect to changes in grid voltage.

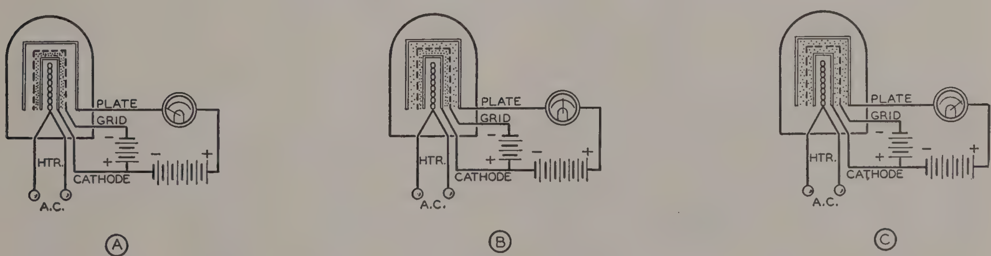


Figure 3.

ANALOGY OF THE ACTION OF THE GRID IN A TRIODE.

(A) shows the tube with cutoff bias on the grid. Note that all the electrons emitted by the cathode remain inside the grid mesh. (B) shows the same tube with an intermediate value of bias on the grid. Note the medium plate current and the fact that there is a reserve of electrons remaining within the grid mesh. (C) shows the tube with a value of grid bias (positive or negative) which allows virtually all the electrons emitted by the cathode to be attracted to the plate. Saturation plate current is attained in this case.

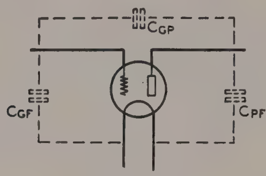


Figure 5.
STATIC INTERELECTRODE CAPACITANCES WITHIN A TRIODE.

This latter point is commonly referred to as the *diode bend* and is caused by the positive voltage of the grid allowing it to rob from the current stream electrons that would normally go to the plate. When the plate voltage is low with respect to that required for full current from the cathode, the diode bend is reached before plate current saturation. When the plate voltage is high, saturation is reached first.

From the above it can be seen that the grid acts as a valve in controlling the electron flow from the cathode to the plate. As long as the grid is kept negative with respect to the cathode, only an extremely small amount of grid energy is required to control a comparatively large amount of plate power. Even if the grid is operated in the positive region a portion of the time, so that it will draw current, the grid energy requirements are still very much less than the energy controlled in the plate circuit. It is for this reason that a vacuum tube is commonly called a *valve* in Britain, Australia, and Canada.

Interelectrode Capacitance. In the preceding chapter it was mentioned that two conductors separated by a dielectric form a *condenser,* or that there is *capacitance* between them. Since the electrodes in a vacuum tube are conductors and they are separated by a dielectric, vacuum, there is capacitance between them. Although the interelectrode capacitances are so small as to be of little consequence in audio-frequency work, they are large enough to be of considerable importance when the tubes are operated at radio frequencies.

Figure 5 shows the interelectrode capacitances of a triode as they appear to a circuit in which the tube is operating. The static capacitances are simply as shown in the drawing, but when a vacuum tube is actually operating as an amplifier there is another consideration known as the *Miller effect* which causes the dynamic input capacitance to be different from the static value. The output capacitance of an amplifier is essentially the same as the static value such as is given in the tube tables. The grid-to-plate capacity is also the same as the static value, but since the C_{gp} acts as a small condenser, coupling energy back from the plate to the grid circuit, the dynamic input capacitance is equal to the static value plus an amount determined by the gain of the stage and the grid-to-plate feedback capacity. Expressed as an equation:

$$C_{gf}\ (\text{dynamic}) = C_{gf}\ (\text{static}) + (M+1)C_{gp}$$

where C_{gf} and C_{gp} are as given before, and M is the stage gain.

In addition to the undesirable Miller effect, whereby the input capacity of an amplifier is increased by the grid-to-plate capacity, this C_{gp} can also cause uncontrollable regeneration or oscillation in radio frequency amplifiers. However, all the undesirable effects of the grid-to-plate capacity can be balanced out by means of a neutralizing circuit. These circuits are discussed under *Neutralization* in the chapter devoted to *Transmitter Theory.*

The Tetrode or Screen-Grid Tube. The quest for a simpler and more easily usable method of eliminating the effects of the grid-to-plate capacity of the triode led to the development of the *screen-grid* tube or *tetrode.* When another grid is added between the grid and plate of a vacuum tube the tube is called a tetrode, and because the new grid is called a *screen,* as a result of its screening or shielding action, the tube is often called a screen-grid tube. The interposed screen grid acts as an electrostatic shield between the grid and plate, with the consequence that the grid-to-plate capacity is reduced. Although the screen grid is maintained at a positive voltage with respect to the cathode of the tube, it is maintained at ground potential with respect to r.f. by means of a by-pass condenser of very low reactance at the frequency of operation.

In addition to the shielding effect, the screen grid serves another very useful purpose. Since the screen is maintained at a positive potential, it serves to increase or accelerate the flow of electrons to the plate. There being large openings in the screen mesh, most of the electrons pass through it and on to the plate. Due also to the screen, the plate current is largely independent of plate voltage, thus making for high amplification. When the screen voltage is held at a constant value, it is possible to make large changes in plate voltage without appreciably affecting the plate current.

Secondary Emission; Pentodes. When the electrons from the cathode approach the plate with sufficient velocity, they dislodge electrons upon striking the plate. This effect of *bombarding* the plate with high velocity electrons, with the consequent dislodgement of other electrons from the plate, is known as *secondary emission.* This effect can cause no

particular difficulty in a triode tube because the secondary electrons so emitted are eventually attracted back to the plate. In the screen-grid tube, however, the screen is close to the plate and is maintained at a positive potential. Thus, the screen will attract these electrons that have been knocked from the plate, particularly when the plate voltage falls to a lower value than the screen voltage, with the result that the plate current is lowered and the amplification is decreased.

This effect is eliminated when still another element is added between the screen and plate. This additional element is called a *suppressor,* and tubes in which it is used are called *pentodes.* The suppressor grid is sometimes connected to cathode within the tube, sometimes it is brought out to a connecting pin on the tube base, but in any case it is established negative with respect to the minimum plate voltage. The secondary electrons that would travel to the screen if there were no suppressor are diverted back to the plate. The plate current is, therefore, not reduced and the amplification possibilities are increased.

Pentodes for radio applications are designed so that the suppressor increases the limits to which the plate voltage may swing; therefore the consequent power output and gain can be very great. Pentodes for radio-frequency service function in such a manner that the suppressor allows high voltage gain, at the same time permitting fairly high gain at low plate voltage. This holds true even if the plate voltage is the same or slightly lower than the screen voltage.

Beam Power Tubes. A beam power tube makes use of a new method for suppressing secondary emission. In this tube there are four electrodes: a cathode, a grid, a screen, and a plate, so spaced and placed that secondary emission from the plate is suppressed without actual power. Because of the manner in which the electrodes are spaced, the electrons which travel to the plate are slowed down when the plate voltage is low, almost to zero velocity in a certain region between screen and plate. For this reason the electrons form a stationary cloud, a *space charge.* The effect of this space charge is to repel secondary electrons emitted from the plate and thus cause them to return to the plate. In this way, secondary emission is suppressed.

Another feature of the beam power tube is the low current drawn by the screen. The screen and the grid are spiral wires wound so that each turn in the screen is shaded from the cathode by a grid turn. This alignment of the screen and the grid causes the electrons to travel in sheets between the turns of the screen so that very few of them strike the screen itself. Because of the effective suppressor action provided by the space charge, and because of the low current drawn by the screen, the beam power tube has the advantages of high power output, high power sensitivity, and high efficiency. The 6L6 is such a beam power tube, designed for use in the power amplifier stages of receivers and speech amplifiers or modulators. Larger tubes employing the beam-power principle are being made by various manufacturers for use in the radio-frequency stages of transmitters. These tubes feature extremely high power sensitivity (a very small amount of driving power is required for a large output), good plate efficiency, and freedom from the requirement for neutralization. Notable among these transmitting beam power tubes are the T21 of Taylor, the 807, 814, and 813 of RCA and G.E., and the HY-65, HY-67, and HY-69 of Hytron.

Television Amplifier Pentodes. There was a need in television work, where extremely wide bands of frequencies must be passed by an amplifier, for vacuum tubes which would give extremely high amplification and still have comparatively low plate impedance and shunt capacitances. This need led to the development of the 1851, 6AB7, 6AC7, 1231, etc.—all of which answer this requirement with slight individual variations. Through the use of a large cathode and a very fine mesh grid spaced very close to the cathode, it has been possible to obtain in these pentodes amplification factors of 6000 and above with transconductances of 5000 to 12,000. The true significance of these figures can be grasped after the material in the latter part of this chapter has been studied.

Pentagrid Converters. A pentagrid converter is a multiple grid tube so designed that the functions of superheterodyne oscillator and mixer are combined in one tube. One of the principal advantages of this type of tube in superheterodyne circuits is that the coupling between oscillator and mixer is automatically accomplished; the oscillator elements effectively modulate the electron stream and, in so doing, the conversion conductance is high. The principal disadvantage of these tubes lies in the fact that they are not particularly suited for operation at frequencies much above 20 Mc. because of difficulties encountered in the oscillator section.

Special Purpose Mixer Tubes. Notable among the special purpose multiple grid tubes is the 6L7 heptode, used principally as a mixer in superheterodyne circuits. This tube has *five* grids: control grid, screens, suppressor and special injection grid for oscillator input. Oscillator coupling to control grid and screen grid circuits of ordinary pentodes is effective as far

as mixing is concerned, but has the disadvantage of considerable interaction between oscillator and mixer.

The 6L7 has a special *injection grid* so placed that it has reasonable effect on the electron stream without the disadvantage of interaction between the screen and control grid. The principal disadvantage is that it requires fairly high oscillator input in order to realize its high conversion conductance. It may also be used as an r.f. pentode amplifier.

The 6J8G and 6K8 are two tubes specifically designed for converter service. They consist of a heptode mixer unit and a triode unit in the same envelope, internally connected to provide the proper injection for conversion work. While both tubes function as a triode oscillator feeding a heptode mixer, the method of injection is different in the two tubes. In the 6J8G, the control grid of the oscillator is connected internally to a special shielded injector grid in the heptode section. In the 6K8, the number one grid of the heptode is connected internally to the control grid of the oscillator triode.

Single-Ended Tubes. From the introduction of the screen-grid tube to the present time, it has been standard practice to bring the control grid (or the no. 1 grid as it is called) of all pentodes and tetrodes designed for radio frequency amplifier use in receivers through the *top* of the envelope. This practice was started because it was much easier to shield the input from the output circuit when one was at the top and the other at the bottom of the envelope. This was true both of the elements and of their associated circuits.

With the introduction of the octal-based metal tube it became feasible to design and manufacture high-gain r.f. amplifier and mixer tubes with all the terminals brought out the base. The metal envelope gives excellent shielding of the elements from external fields, and through the use of a small additional shield inside the locating pin of the octal socket, the diametrically opposite grid and plate pins of the tubes are well shielded from each other. A more or less complete line of tubes for ordinary receiving service has been made available in the single-ended type. The type numbers of these tubes contain an *S* between the filament voltage and the rating classification letter as: 6SA7, 6SQ7, 12SK7, etc.

Another type of single-ended tube which has come into prominent usage is the *loctal* group. These loctal tubes are all glass with a metal base and metal locating pin; the tube prongs extend through the bottom of the glass envelope and make direct connection to the elements of the tube. Due to the shortness and directness of the leads to the elements, loctal tubes are generally conceded to be the most satisfactory type for high-frequency work. A quite complete line for all ordinary receiving purposes is now being manufactured in the loctal type. The distinguishing feature of the loctal tube numbers is the fact that these numbers start out with a 7 or a 14 instead of the 6 or 12 used in conventional receiving types, as: 7A7, 7C5, 14A7, etc. The heater voltage ratings, however, are 6.3 or 12.6 volts as in the other conventional types. Loctal tubes are also made in the 1.4-volt series; these have a characteristic number beginning with 1L such as: 1LA4, 1LA6, etc.

Dual Tubes. Some of the commonly known vacuum tubes are in reality two tubes in one, i.e., in a single glass or metal envelope. Twin triodes, such as the types 53, 6A6, 6SC7, and 6N7 are examples. A disadvantage of these twin-triode tubes for certain applications is the fact that the cathodes of both tubes are brought out to the same base pin.

Of a different nature are the 6H6 and 7A6 twin diodes and the 6F8G, 6SN7-GT, 7F7, and 6C8G twin triodes. The cathodes of each of these tubes are brought to a separate base pin on the socket, thus making them true twin tubes. Other types combine the functions of a double diode and either low or high μ triode in the same envelope, as well as a similar combination with a pentode instead of a triode. Still other types combine a pentode and a triode, a pentode and a power supply rectifier, and electron-ray indicating tubes (magic eyes) with their self-contained triode d.c. voltage amplifier.

Manufacturer's Tube Manuals. The larger tube manufacturers offer at a nominal cost, tube manuals which are very complete and give much valuable data which, because of space limitations, cannot be included in this handbook. Those especially interested in vacuum tubes are urged to purchase one of these books as a supplementary reference.

APPLICATION AND OPERATION OF THE VACUUM TUBE

The preceding section of this chapter has been devoted to the general theory of vacuum tubes and to the various forms in which they commonly appear. The succeeding section will be devoted to the application of the characteristics and abilities of the vacuum tube to the problems of amplification, oscillation, rectification, detection, frequency conversion, and electrical measurements.

The Vacuum Tube as an Amplifier

The ability of a grid of a vacuum tube to control large amounts of plate power with a small amount of input energy allows the vacuum tube to be used as an amplifier. It is the ability of the vacuum tube to amplify an extremely small amount of energy up to almost any amount without change in anything except amplitude which makes the vacuum tube such an extremely useful adjunct to modern industry and communication.

The most important considerations of a vacuum tube, aside from its power handling ability (which will be treated later on), are amplification factor, plate resistance, and mutual conductance or transconductance.

Amplification Factor or Mu. The amplification factor or mu (μ) of a vacuum tube is the ratio of a change in plate voltage to a change in grid voltage, either of which will cause the same change in plate current. Expressed as a differential equation:

$$\mu = -\frac{dE_p}{dE_g} \qquad I_p = \text{constant}$$

The μ can be determined experimentally by making a slight change in the plate voltage, thus slightly changing the plate current. The plate current is then returned to its original value by a change in grid voltage. The ratio of the increment in plate voltage to the increment in grid voltage is the μ of the tube. The foregoing assumes that the experiment is conducted on the basis of rated voltages as shown in the manufacturer's tube tables.

Plate Resistance. The plate resistance of a vacuum tube is the ratio of a change in plate voltage to the change in plate current which the voltage change produces. To be accurate, the changes should be very small with respect to the operating values. Expressed as an equation:

$$R_p = \frac{dE_p}{dI_p}$$

The plate resistance can also be determined by the experiment mentioned above. By noting the change in plate current as it occurs when the plate voltage is changed, and by dividing the latter by the former, the plate resistance can then be determined. Plate resistance is expressed in ohms.

Transconductance. The mutual conductance, also referred to as *transconductance,* is the ratio of the amplification factor (μ) to the plate resistance:

$$G_M = \frac{\mu}{R_p} = \frac{\dfrac{dE_p}{dE_g}}{\dfrac{dE_p}{dI_p}} = \frac{dI_p}{dE_g}$$

Transconductance is most commonly expressed in micro-reciprocal-ohms or *micromhos*. However, since transconductance expresses change in plate current as a function of a change in grid voltage, a tube is often said to have a transconductance of so many milliamperes-per-volt. If the transconductance in milliamperes-per-volt is multiplied by 1000 it will then be expressed in micromhos. Thus the transconductance of a 6A3 could be called either 5.25 ma./volt or 5250 micromhos.

The transconductance is probably the most important single characteristic of a vacuum tube. It is often called the figure of merit because the G_M is an excellent indication of the effectiveness of a tube as an amplifier and of its power sensitivity—the greater the transconductance, the greater will be the gain of an r.f. amplifier, and the greater will be the power output with a given grid voltage of a power audio amplifier.

Gain per Stage. When a vacuum tube is used as a resistance-coupled audio amplifier, it is important to know in advance just how much gain will be obtained from a particular stage. The stage gain of a large number of common vacuum tubes under various circuit conditions is given in the *RCA Receiving Tube Manual* (25¢ from RCA) and in other vacuum-tube manuals. However, when it is desired to know what a specific tube will do under certain specified operating conditions, the following two formulas will be of assistance—in either case they will indicate the gain in voltage to be expected from a stage at a medium audio frequency in the vicinity of 1000 cycles. The stage gain at extremely high and low frequencies will be determined by the values of resistance and capacitance making up the circuit.

$$\textbf{Gain, triode amplifier} = \frac{\mu R_L}{R_p + R_L}$$

where: $_L$ is the amplification factor of the tube
R_L is the plate load resistance of the stage
R_p is the plate resistance of the tube.

$$\textbf{Gain, pentode amplifier} = G_M\, R_L$$

where: G_M is the tube transconductance in *mhos* (micromhos/10^6)
R_L is the load resistance of the stage and where the plate resistance of the tube is large compared to the load resistance.

As a practical example of the method of determining the gain of a triode amplifier, suppose we take the case of a 6F5 tube with a plate resistance of 66,000 ohms and an amplification factor of 100 operating into a load resistance of 50,000 ohms. The voltage amplification of the stage as calculated from the above equation would be:

$$\frac{100 \times 50{,}000}{50{,}000 + 66{,}000} = 43$$

From the foregoing it is seen that an input of 1 volt to the grid of the tube will give an output of 43 volts (a.c.).

The calculation of the approximate gain of a resistance-coupled pentode audio stage is even more simple. Suppose the amplifier tube is a 6SJ7 with a transconductance of 1600 micromhos (from the tube tables). This tube's transconductance in *mhos* would be (taking off 6 decimal places) 0.0016 mhos. If the load resistance of the tube is 100,000 ohms, the gain would be (pointing ahead 5 places to multiply 0.0016 by 100,000) 160.

Audio-Frequency Amplifiers

Amplifiers designed to operate at a low level at radio, intermediate, and audio frequencies are almost invariably of the class A type. Higher level audio amplifiers can be of the class A, class AB, or class B type; these classifications and their considerations will be considered first. The class B and class C amplifiers as used for medium and high-level radio-frequency work will be considered under *Radio-Frequency Amplifiers*.

The Class A Amplifier. A class A amplifier is, by definition, *an amplifier in which the grid bias and alternating grid voltages are such that plate current in a specific tube flows at all times*. The output waveform from a class A amplifier is a faithful reproduction of the exciting a.c. voltage upon the grid. For the above conditions to be the case it is necessary that the grid bias, or the operating point, of the amplifier be chosen with care to allow maximum output with minimum distortion.

Figure 6 shows the operating characteristic of a typical triode vacuum tube. It will be noticed that the curve of plate current with varying grid voltage is quite linear within certain limits—outside these limits it is no longer a straight line. For an amplifier to be able to put out a voltage waveform which is a faithful reproduction of the input waveform, it is necessary that the range over which the grid voltage will be varied shall give a linear variation in plate current. Also, a class A amplifier must not draw grid current; so the operating point must be midway between the point of zero grid bias and the point on the operating characteristic where the curvature becomes noticeable. Such a point has been chosen graphically in Figure 6.

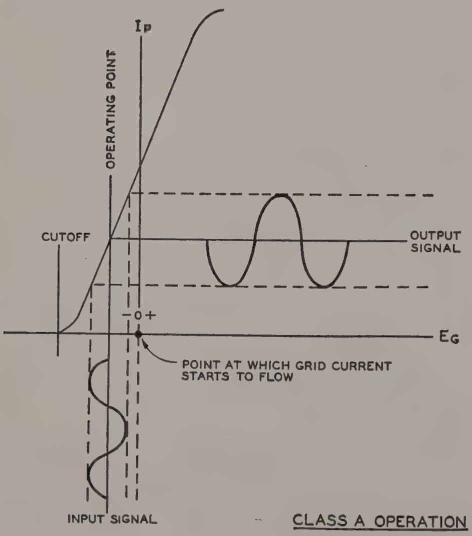
When the grid bias is varied around this operating point, the fluctuation in grid potential results in a corresponding fluctuation in plate current. When this current flows through a suitable load device, it produces a varying voltage drop which is a replica of the original input voltage, although considerably greater in amplitude.

Should the signal voltage on the grid be permitted to go too far negative, the negative half cycle in the plate output will not be the same as the positive half cycle. In other words, the output wave shape will not be a duplicate of the input, and *distortion* in the output will therefore result. The fundamental property of class A amplification is that the bias voltage and input signal level must not advance beyond the point of zero grid potential; otherwise, the grid itself will become positive. Electrons will then flow into the grid and through its external circuit in much the same manner as if the grid were actually the plate. The result of such a flow of grid current is a lowering of the input impedance of the tube so that power is required to drive it.

Since class A amplifiers are never designed to draw grid current, they do not realize the optimum capabilities of any individual tube.

Inspection of the operating characteristic of Figure 6 reveals that there is a long stretch of

Figure 6.



linear characteristic far into the positive grid region. As only the small portion of the operating characteristic below the zero grid bias line can be used, the plate circuit efficiency of a class A amplifier is low. However, they are used because they have very little or negligible distortion and, since only an infinitesimal amount of power is required on the grid, a large amount of power amplification may be obtained. Low-level audio and radio frequency amplifying stages in receivers and audio amplifiers are invariably operated class A. The correct values of bias for the operation of tubes as class A amplifiers are given in the *Tube Tables.*

The Class AB Amplifier. A class AB amplifier is one in which *the grid bias and alternating grid voltages are such that plate current in a specific tube flows for appreciably more than half but less than the entire electrical cycle when delivering maximum output.*

In a class AB amplifier, the fixed grid bias is made higher than would be the case for a push-pull class A amplifier. The resting plate current is thereby reduced and higher values of plate voltage can be used without exceeding the rated plate dissipation of the tube. The result is an increase in power output.

Class AB amplifiers can be subdivided into class AB_1 and class AB_2. There is no flow of grid current in a class AB_1 amplifier; that is, the peak signal voltage applied to each grid does not exceed the negative grid bias voltage. In a class AB_2 amplifier, the grid signal is greater than the bias voltage on the peaks, and grid current flows.

The class AB amplifier should be operated in push-pull if distortion is to be held to a minimum. Class AB_2 will furnish more power output for a given pair of tubes than will class AB_1. The grids of a class AB_2 amplifier draw current, which calls for a power driver stage.

The Class B Amplifier. A class B amplifier is one in which *the grid bias is approximately equal to the cutoff value so that the plate current is very low (almost zero) when no exciting grid voltage is applied and so that plate current in a specific tube flows for approximately one half of each cycle when an alternating grid voltage is applied.*

A class B *audio* amplifier always operates with two tubes in push-pull. The bias voltage is increased to the point where but very little plate current flows. This point is called the *cutoff point.* When the grids are fed with voltage 180 degrees out of phase, that is, one grid swinging in a positive direction and the other in a negative direction, the two tubes will alternately supply current to the load.

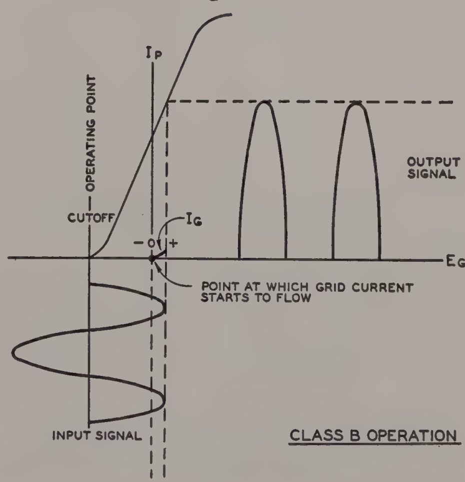
When the grid of tube no. 1 swings in a positive direction, plate current flows in this tube. During this process, grid no. 2 swings negatively beyond the point of cutoff; hence, no current flows in tube no. 2. On the other half-cycle, tube no. 1 is idle, and tube no. 2 furnishes current. Each tube operates on one-half cycle of the input voltage so that the complete input wave is reproduced in the plate circuit. Since the plate current rests at a very low value when no signal is applied, the plate efficiency is considerably higher than in a class A amplifier.

There is a much higher, steady value of plate current flow in a class A amplifier, regardless of whether or not a signal is present. The average plate dissipation or plate loss is much greater than in a class B amplifier of the same power output capability.

For the reason that the plate current rises from a very low to a very high peak value on input swings in a class B audio amplifier, the demands upon the power supply are quite severe; a power supply for class B amplifier service must have good *regulation.* A high-capacity output condenser must be used in the filter circuit to give sufficient storage to supply power for the stronger audio peaks, and a choke-input filter system is required for good regulation.

Load Impedance for Amplifiers. The plate current in an amplifier increases and decreases in proportion to the value of applied input signal. If useful power is to be realized from such an amplifier, the plate circuit must be terminated in a suitable resistance or impedance across which the power can be developed. When increasing and decreasing plate current flows through a resistor or impedance,

Figure 7.



the voltage drop across this load will constantly change because the plate current is constantly changing. The actual value of voltage on the plate will vary in accordance with the IZ drop across the load, even though a steady value of direct current may be applied to the load impedance; hence, for an alternating voltage on the grid of the tube, there will be a constant change in the voltage at the anode.

The static characteristic curves give an indication of the performance of the tube for only one value of plate voltage. If the plate voltage is changed, the characteristic curve will shift. This sequence of change can be plotted in a form that permits a determination of tube performance; it is customary to plot the plate current for a series of permissible values of plate voltage at some fixed value of grid voltage.

The process is repeated for a sufficient number of grid voltage values in order that adequate data will be available. A group or family of plate voltage—plate current curves, each for a different grid potential, makes possible the calculation of the correct load impedance for the tube. Dynamic characteristics include curves for variations in amplification factor, plate resistance, transconductance and detector characteristics.

The correct value of load impedance for a rated power output is always specified by the tube manufacturer. The plate coupling device generally reflects this impedance to the tube. This subject is treated under *Impedance Matching,* Chapter 2.

Tubes in Parallel and Push-Pull. Two or more tubes can be connected in parallel in order to secure greater power output; two tubes in parallel will give approximately twice the output of a single tube. Since the plate resistances of the two tubes are in parallel, the required load impedance will be half that for a single tube.

When power is to be increased by the use of two tubes, it is generally advisable to connect them in push-pull; in this connection the power output is doubled and the *harmonic content,* or *distortion,* is reduced. The input voltage applied to the grids of two tubes is 180 degrees out of phase, the voltage usually being secured from a center-tapped secondary winding with the center tap connected to the source of bias and the outer ends of the winding connected to each grid. The plates are similarly fed into a center-tapped winding and plate voltage is introduced at the center tap. The signal voltage supplied to one grid must always swing in a positive direction when the other grid swings negatively. The result is an increase in plate current in one tube with a decrease in plate current in the other at any given instant; one tube *pushes* as the other *pulls;* hence the term: *push-pull.*

Distortion in Audio Amplifiers. Distortion exists when the output wave shape of an amplifier differs from the shape of the input voltage wave. There are three main types of distortion which can exist in an audio amplifier. These are: frequency distortion, where the gain of the amplifier is not the same for all frequencies which are to be passed; non-linear distortion, which results in cross modulation of the various audio frequencies fed into the amplifier and which also results in the production of harmonics of these tones; and phase distortion, which is the result of the amplifier's having different delay characteristics for various audio frequencies.

Frequency distortion can be kept to a minimum through the use of high-quality audio transformers wherever transformers are needed and through the use of the proper values of coupling and by-pass condensers and feed resistors in the resistance coupled stages. Careful choice of components can result in an audio amplifier which is "flat within 1 db from 25 to 15,000 cycles"; such an amplifier would give high quality reproduction, provided non-linear and phase distortion were also at a minimum.

Non-linear distortion is usually caused by the overloading of some vacuum tube within the amplifier—usually the output stage. The presence of non-linear distortion is usually expressed by the rating of the amplifier at a certain percentage of r.m.s. harmonic distortion at a certain amount of output power. The amount of non-linear distortion almost invariably increases with increasing power output from the amplifier. Non-linear distortion is peculiar in that it always results in the production of frequencies in the output waveshape of the amplifier which were not present in the input. Since these spurious frequencies resulting from non-linear distortion are mainly in the form of *harmonics* of the input frequency (integral multiples: second harmonic, twice frequency; third harmonic, three times frequency, etc.), non-linear distortion is usually called *harmonic distortion.*

The presence of strong harmonics in an audio frequency amplifier gives rise to speech and music distortion which is plainly apparent to the human ear. Triode amplifiers give rise to distortion which is mainly second harmonic, pentodes and tetrodes give rise to more third harmonic distortion than second, while a balanced push-pull amplifier produces only odd harmonic distortion (third, fifth, seventh, etc.). Third harmonic distortion is much more apparent to the ear than second harmonic. Since the harmonic distortion naturally falls in the higher frequency region of reproduction (har-

monics are multiples of their generating wave frequencies), the upper frequency limit of the reproducing system determines the maximum amount of harmonic distortion which can be permitted. In a conventional reproducing system such as is found in the average good quality receiver, the value of 5 per cent is generally accepted as the maximum permissible total harmonic distortion; of this total value not more than 2 per cent should be attributable to third and higher order harmonics for good reproduced quality.

Phase distortion is not generally considered to be of great importance in a single audio or speech amplifier. However, when a large number of audio amplifiers are cascaded, as in long wire line repeater amplifiers, the cumulative phase distortion can become serious; properly designed delay circuits and other measures are taken to correct the normal delay under these conditions. Phase distortion must also be kept to a minimum for proper operation of television video amplifiers and of audio amplifiers into which degenerative feedback is to be incorporated.

Voltage and Power Amplification. Practically all amplifiers can be divided into two classifications: *voltage amplifiers* and *power amplifiers*. In a voltage amplifier, it is desirable to increase the voltage to a maximum possible value, consistent with allowable distortion. The tube is not required to furnish *power* because the succeeding tube is always biased to the point where no grid current flows. The selection of a tube for voltage amplifier service depends upon the voltage amplification it must provide, upon the load that is to be used and upon the available value of plate voltage. The varying signal current in the plate circuit of a voltage amplifier is employed in the plate load solely in the production of *voltage* to be applied to the grid of the following stage. The plate voltage is always relatively high, the plate current small.

A *power amplifier,* in contrast, must be capable of supplying a heavy current into a load impedance that usually lies between 2000 and 20,000 ohms. Power amplifiers normally furnish excitation to power-consuming devices such as loud-speakers and modulated class C amplifiers. They also serve as drivers for other larger amplifier stages whose grids require power from the preceding stage. Power amplifiers are common in transmitters.

The difference between the plate power input and output is dissipated in the tube in the form of heat, and is known as the *plate dissipation.* Tubes for power amplifier service have larger plates and heavier filaments than those for a voltage amplifier. High-power audio circuits for commercial broadcast transmitters call for tubes of such proportions that it becomes necessary to cool their plates by means of water or forced-air cooling systems.

Interstage Coupling. Common methods of coupling one stage to another in an audio amplifier are shown in Figure 8.

Transformer coupling for a single-ended stage is shown in A; coupling to a *push-pull* stage in B; *resistance coupling* in C; *impedance coupling* in D. A combination *impedance-transformer* coupling system is shown in

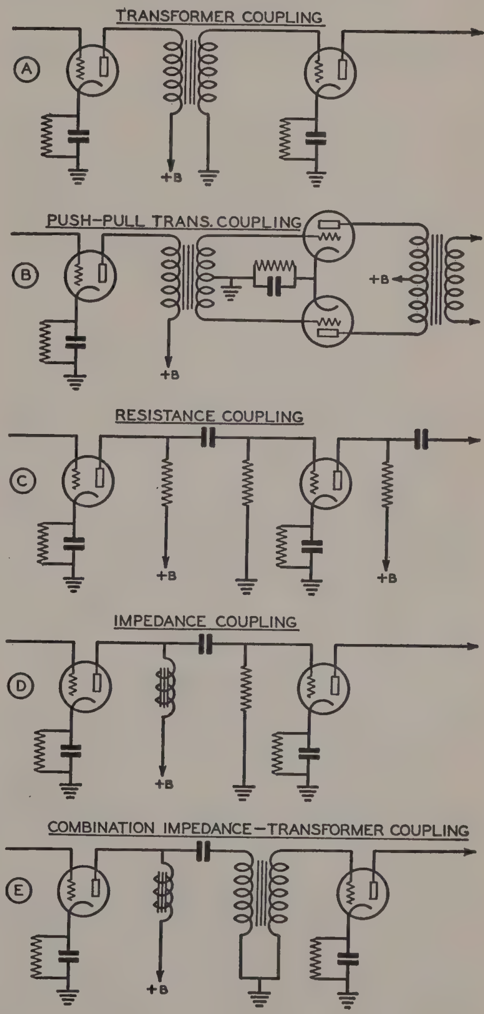


Figure 8.
FIVE COMMON METHODS OF AUDIO-FREQUENCY INTERSTAGE COUPLING.

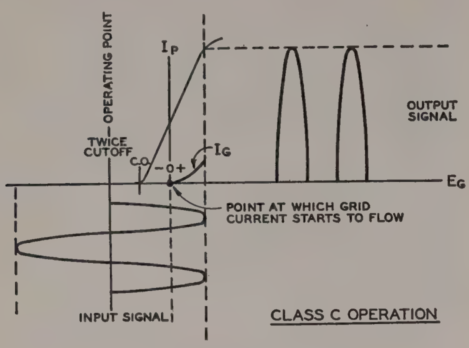


Figure 9.

E; this arrangement is generally chosen for high permeability audio transformers of small size and where it is necessary to prevent the plate current from flowing through the transformer primary. The plate circuit in the latter is *shunt-fed.* A resistor of appropriate value is often substituted for the impedance in the circuit shown in E.

Radio-Frequency Amplifiers

Radio-frequency amplifiers, as used in transmitters, invariably fall into the "power" classification. Also, since they operate into sharply tuned tank circuits which tend to take out irregularities in the plate current waveform and give a comparatively pure sine-wave output, more efficient conditions of operation may be used than for an audio amplifier in which the output waveform must be the same as the input over a wide band of frequencies. The class B and class C r.f. amplifiers fall into this grouping.

The Class B R.F. Amplifier. The definition of a class B r.f. amplifier is the same as that of a class B amplifier for audio use. However, the r.f. amplifier operates into a tuned circuit and covers only a very small range of frequencies, while the audio type works into an untuned load and may cover a range of 500 or 1000 to 1 in frequency.

Class B radio-frequency amplifiers are used primarily as *linear amplifiers* whose function is to increase the output from a modulated class C stage. The bias is adjusted to the cutoff value. In a single-ended stage, the r.f. plate current flows on alternate half cycles. The power output in class B r.f. amplifiers is proportional to the square of the grid excitation voltage. The grid voltage excitation is doubled in a linear amplifier at 100 per cent modulation, the grid excitation voltage being supplied by the modulated stage; hence, the power output on modulation peaks in a linear stage is increased four times in value. In spite of the fact that power is supplied to the tank circuit only on alternate half cycles by a single-ended class B r.f. amplifier, the flywheel effect of the tuned tank circuit supplies the missing half cycle of radio frequency, and the complete waveform is reproduced in the output to the antenna.

The Class C R.F. Amplifier. A class C amplifier is defined as an amplifier in which the grid bias is appreciably greater than the cutoff value so that the plate current in each tube is zero when no alternating grid voltage is applied, and so that plate current in a specific tube flows for appreciably less than one half of each cycle when an alternating grid voltage is applied.

Angle of Plate Current Flow. The class C amplifier differs from others in that the bias voltage is increased to a point well beyond cutoff. When a tube is biased to cutoff, as in a class B amplifier, it draws plate current for a half cycle or 180°. As this point of operation is carried beyond cutoff, that is, when the grid bias becomes more negative, the angle of plate current flow decreases. Under normal conditions, the optimum value for class C amplifier operation is approximately 120°. The plate current is at zero value during the first 30° because the grid voltage is still approaching cutoff. From 30° to 90°, the grid voltage has advanced beyond cutoff and swings to a maximum in a region which allows plate current to flow. From 90° to 150°, the grid voltage returns to cutoff, and the plate current decreases to zero. From 150° to 180°, no plate current flows, since the grid voltage is then beyond cutoff.

The plate current in a class C amplifier flows in pulses of high amplitude, but of short duration. Efficiencies up to 75 per cent are realized under these conditions. It is possible to convert nearly all of the plate input power into r.f. output power (approximately 90 per cent efficiency) by increasing the excitation, plate voltage, and bias to extreme values.

Linearity of Class C Amplifiers. The r.f. plate current is proportional to the plate voltage; hence the power output is proportional to the square of the plate voltage. Class C amplifiers are invariably used for plate modulation because of their high efficiency and because they reflect a pure resistance load into the modulator. The plate voltage of the class C stage is doubled on the peaks at 100 per cent modulation; since the plate current is also doubled, the power output at this point is consequently increased four times.

Figure 9 illustrates graphically the operation of a class C amplifier with twice cut-off bias and with the peak grid swing of such a value as just to approach the diode bend in the plate characteristic. When the excitation voltage is increased beyond this point, the plate current waveform will have a dip at the crest due to the taking of electrons from the plate current stream by the grid on its highly positive peaks.

The Vacuum Tube as an Oscillator

The ability of an amplifier tube to control power enables it to function as an oscillator or a generator of alternating current in a suitable circuit. When part of the amplified output is coupled back into the input circuit, sustained oscillations will be generated provided the input voltage to the grid is of the proper magnitude and phase with respect to the plate.

The voltage that is fed back and applied to the grid must be 180° out of phase with the voltage across the load impedance in the plate circuit. The voltage swings are of a frequency depending upon circuit constants.

If a parallel resonant circuit consisting of an inductance and capacitance is inserted in series with the plate circuit of an amplifier tube and a connection is made so that part of the potential drop is impressed 180° out of phase on the grid of the same tube, amplification of the potential across the L/C circuit will result. The potential would increase to an unrestricted value were it not for the limited plate voltage and the limited range of linearity of the tube characteristic, which causes a reversal of the process after a certain point is reached. The rate of reversal is determined by the time constant or resonant frequency of the tank circuit.

The frequency range of an oscillator can be made very great; thus, by varying the circuit constants, oscillations from a few cycles per second up to many millions can be generated. A number of different types of oscillators are treated in detail under the section devoted to *Transmitter Theory*.

The Vacuum Tube as a Rectifier

It was stated at the first of this chapter that when the potential of the plate of a two-element vacuum tube or diode is made positive with respect to the cathode, electrons emitted by the cathode will be attracted to the plate and a current will flow in the external circuit that returns the electrons to the cathode. If, on the other hand, the plate is made negative with respect to the cathode the electron flow in the external circuit will cease, due to the repulsion of the electronic stream within the tube back

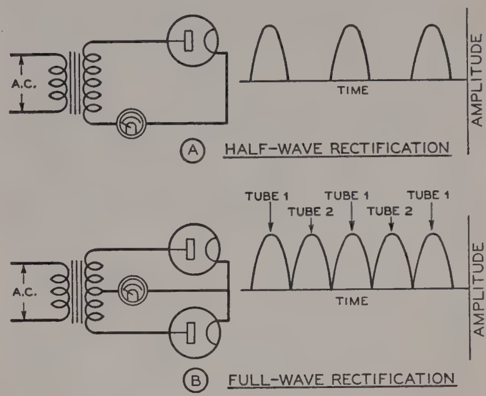


Figure 10.

to the cathode. From this is derived a valuable property, namely, the ability of a vacuum tube to pass current in one direction only and hence to function as a *rectifier* or a device to convert alternating current into pulsating d.c.

The Half-Wave Rectifier. Figure 10A shows a half-wave rectifier circuit. For convenience of explanation, a conventional power rectifier has been chosen, although the same diagram and explanation would apply to diode rectification as employed in the detector circuits of many receivers.

When a sine-wave voltage is induced in the secondary of the transformer, the rectifier plate is made alternately positive and negative as the polarity of the alternating current changes. Electrons are attracted to the plate from the cathode when the plate is positive, and current then flows in the external circuit. On the succeeding half cycle, the plate becomes negative with respect to the cathode, and no current flows. Thus, there will be an interval before the succeeding half cycle occurs when the plate again becomes positive. Under these conditions, plate current once more begins to flow, and there is another pulsation in the output circuit.

For the reason that one half of the complete wave is absent in the output, the result is what is known as *half-wave rectification*. The output power is the average value of these pulsations; it will, therefore, be of a low value because of the interval between pulsations.

Full-Wave Rectification. In a *full-wave circuit* (Figure 10B), the plate of one tube is positive when the other plate is negative; although the current changes its polarity, one of the plates is always positive. One tube, therefore, operates effectively on each half

cycle, but the output current is in the same direction. In this type of circuit the rectification is complete and there is no gap between plate current pulsations. This output is known as *rectified a.c.* or *pulsating d.c.*

Mercury Vapor Rectifiers. If a two-element electron tube is evacuated and then filled with a gas such as mercury vapor, its characteristics and performance will differ radically from those of an ordinary high-vacuum diode tube.

The principle upon which the operation of a gas-filled rectifier depends is known as the phenomenon of gaseous ionization, which was discussed under *Fundamental Theory*. Investigation has shown that the electrons emitted by a hot cathode in a mercury-vapor tube are accelerated toward the anode (plate) with great velocity. These electrons move in the electrical field between the hot cathode and the anode. In this space they collide with the mercury-vapor molecules which are present.

If the moving electrons attain a velocity so great as to enable them to break through a potential difference of more than 10.4 volts (for mercury), they will literally knock the electrons out of the atoms with which they collide.

As more and more atoms are broken up by collision with electrons, the mercury vapor within the tube becomes *ionized* and transmits a considerable amount of current. The ions are repelled from the anode when it is positive; they are then attracted to the cathode, thus tending to neutralize the negative space charge as long as saturation current is not drawn. This effect neutralizes the negative space charge to such a degree that the voltage drop across the tube is reduced to a very low and constant value. Furthermore, a considerable reduction in heating of the diode plate, as well as an improvement in the voltage regulation of the load current, is achieved. The efficiency of rectification is thereby increased because the voltage drop across any rectifier tube represents a waste of power.

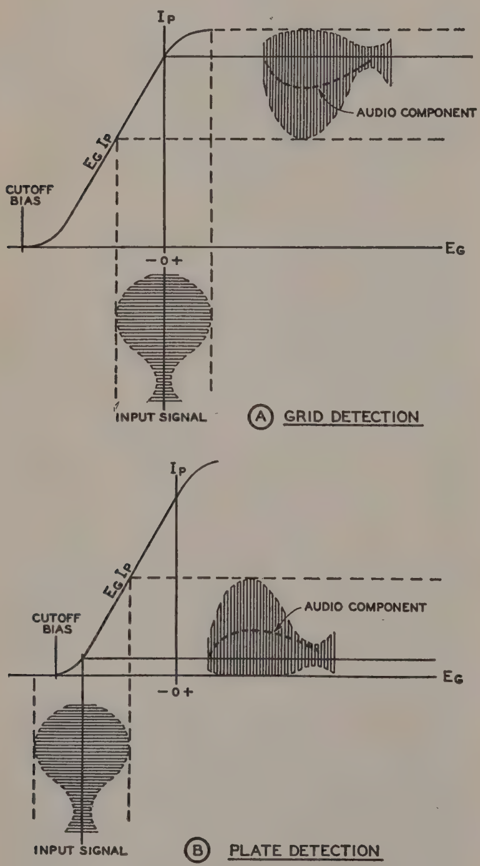


Figure 11.
ILLUSTRATING DETECTOR OPERATION IN UPPER AND LOWER-BEND PORTION OF THE CHARACTERISTIC CURVE.

Detection or Demodulation

Detection is the process by which the audio component is separated from the modulated radio-frequency signal carrier at the receiver. Detection always involves either rectification or nonlinear amplification of an alternating current.

Two general types of amplifying detectors are used in radio circuits.

The Plate Detector. The plate detector or *bias detector* (sometimes called a *power detector*) amplifies the radio-frequency wave and then rectifies it, and passes the resultant audio signal component to the succeeding audio amplifier. The detector operates on the lower bend in the plate current characteristic, because it is biased close to the cutoff point and therefore could be called a single-ended class B amplifier. The plate current is quite low in the absence of a signal, and the audio component is evidenced by an increase in the average unmodulated plate current. See Figure 11.

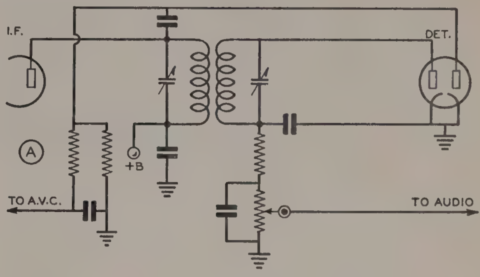
The Grid Detector. The grid detector differs from the plate detector in that it rectifies in the grid circuit and then amplifies the resultant audio signal. The only source of grid bias is the grid leak so that the plate current is maximum when no signal is present. This

form of detector operates on the upper or saturated bend of its characteristic curve and the demodulated signal appears as an audio-frequency decrease in the average plate current. However, at *low* plate voltage, most of the rectification takes place as the result of the *curvature* in the grid characteristic. By proper choice of grid leak and plate voltage, distortion can be held to a reasonably small value. In extreme cases the distortion can reach a very high value, particularly when the carrier signal is modulated to a high percentage. In such cases the distortion can reach 25 per cent.

The grid detector will absorb some power from the preceding stage because it draws grid current. It is significant to relate that the higher gain through the grid detector does not necessarily indicate that it is more sensitive. Detector sensitivity is a matter of *rectification efficiency* and amplification, not of amplification alone. Grid leak detectors are often used in regenerative detector circuits because smoother control of *regeneration* is possible than in other forms of plate and bias detectors.

Non-Amplifying Detectors. In addition to the two previous types of amplifying detectors, both of which have a certain inherent amount of harmonic distortion, there are two main types of non-amplifying detectors which have, of late, been more widely used because of their lowered harmonic distortion and other advantages.

The Diode Detector. In this type of detector the input r.f. signal (almost invariably at the intermediate frequency of the receiver) is simply rectified by the diode and the modulation component appears as an alternating voltage, in addition to the d.c. component, across the diode load resistor. This type of detection, although it gives no gain and has a loading effect on the circuit that feeds it, is frequently used in high-quality receivers because of the relatively distortionless detection or demodulation that is obtained. Figure 12A shows a combined detector-a.v.c. rectifier circuit commonly used in high-quality receivers. It will be noticed that a separate diode and rectifier circuit is used to obtain the a.v.c. voltage. This is done to eliminate the *a.c. shunt loading* of the a.v.c. bus upon the detector circuit. If the a.v.c. voltage is taken from the detector diode load resistor, the effect of the a.c. shunt loading of the a.v.c. circuit can be serious enough to cause as high as 25 per cent harmonic distortion of a 100 per cent modulated input signal. However, inexpensive midget receivers in which the high-frequency response is limited, frequently take the a.v.c. voltage from the detector load resistor and rely upon the limited high-frequency response to make the distortion unnoticeable.



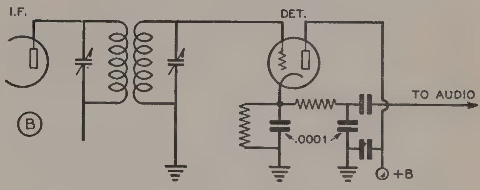


Figure 12.
(A) DIODE DETECTOR WITH SEPARATE A.V.C. RECTIFIER. (B) INFINITE IMPEDANCE DETECTOR.

Certain circuits are available for compensating for the a.c. shunt loading effect of the a.v.c. circuit upon the detector load resistor, but the most satisfactory arrangement is that shown in 12A in which a separate rectifier taking its r.f. voltage from the plate of the last i.f. amplifier is used to supply the a.v.c. voltage. It is also best that the lead marked "to audio" in Figure 12A connect directly to the first audio grid, and thus that diode biasing be used upon this grid. If an additional condenser and potentiometer is used between the diode load resistor and the first audio stage, the shunt loading effect of the additional volume control resistor can be as serious as the a.c. loading of the a.v.c. circuit.

The Infinite Impedance Detector. Figure 12B illustrates this comparatively recently popularized type of detector circuit which has advantages over previous types where distortion-free detection is required. The circuit is essentially the same as that for plate or power-detection, except that the output voltage is taken from the cathode circuit instead of from the plate. This gives the advantage that practically 100 per cent degenerative feedback is incorporated into the circuit with a consequent great reduction in harmonic distortion as compared to the simple plate detector. The circuit gives no loading to the circuit from which it obtains its voltage—hence the name, infinite-impedance detection. Also, due to the 100 per cent degenerative

feedback, the circuit has a gain of one. Essentially the same output voltage will be obtained from this detector as will be obtained from a diode detector.

When automatic volume control is to be used in a receiver which employs an infinite impedance detector, the a.v.c. rectifier circuit shown using the right hand diode of Figure 12A can be used. It is common practice to use a combination tube such as the 6B8 as a combined last i.f. amplifier and a.v.c. rectifier, with a separate tube such as a 6J5 as the infinite impedance detector.

Frequency Converters or Mixers. Another common usage of the vacuum tube is as a frequency changer or mixer tube. This is the operation performed by the first detector or mixer in a super-heterodyne, and consists of changing (most frequently) a particular high-frequency signal (bearing the desired modulation) to a fixed intermediate frequency. In this service, the high-frequency signal and another signal from a local oscillator, whose frequency is either lower or higher than the h.f. signal by an amount equal to the intermediate frequency (the frequency to which it is desired to convert), are fed to appropriate grids of the converter tube. The resultant intermodulation of the two signals in the converter tube produces one frequency which is the sum of the two, and another frequency which is equal to the difference between their frequencies. It is this latter frequency which is selected by the output circuit of the mixer tube, and which is subsequently fed to the intermediate frequency amplifier.

Conversion Conductance. The relative efficiency of a converter tube in changing one frequency to another is called its conversion conductance or transconductance. Recent improvements in mixer tubes have allowed sizeable improvements to be made in the efficiency of mixer stages. With the latest types of mixer tubes it is possible to obtain nearly as much gain from a frequency changing stage as from an amplifier stage with its input and output circuits on the same frequency. Discussion of mixer characteristics will be found in the chapter, *Receiver Theory,* and under the section *Special Purpose Mixer Tubes* earlier in this chapter.

The Vacuum Tube as a Measuring Device. The characteristics of the vacuum tube make it very well suited for use as a measuring device in electrical circuits, especially when no power may be taken from the circuit under measurement. Vacuum tube voltmeters are the most common application of this principle. V.t. voltmeters of the peak-indicating and r.m.s. types will be found in the chapter *Test and Measurement Equipment.*

Particular types of vacuum tube voltmeters utilizing the action of an electron stream upon a fluorescent material to give a visual indication are the electron-ray or "magic-eye" tubes, and the cathode-ray oscilloscope. In the electron-ray tube a small knife whose charge varies with the voltage under measurement (usually the amplified d.c. voltage of an a.v.c. circuit) deflects the electron stream to produce a varying angle of fluorescence on the visible screen at the end of the tube.

In the cathode-ray tube an electron gun consisting of cathode, grid, and accelerating anode or plate (the "electron gun") shoots a fine beam of electrons between two sets of deflecting plates separated by 90° to a fluorescent viewing screen at the end of the tube. One set of deflecting plates is most commonly set up so that it will deflect the stream of electrons back and forth in the horizontal plane. The other set of deflecting plates is oriented so that it will deflect the same stream up and down in the vertical plane. The practical design, construction, and application of the cathode-ray oscilloscope to the problems of the amateur station is covered in the *Test and Measurement Equipment* chapter.

Radio Receiver Theory

A RADIO receiver may be defined as a device for reproducing in the form of useful output the intelligence conveyed by radio waves applied to it. Usually an antenna is a necessary adjunct to the receiver. The antenna will not be discussed in this chapter, however, as the function and design of antennas is thoroughly covered in a later section of this book.

Detection. All receivers use some sort of detector to make audible the intelligence impressed on the radiated carrier wave at the transmitter. The process of impressing the intelligence on the carrier wave is known as modulation, and as the detector separates this modulation from the carrier, it is often known as a *demodulator*. One of the simplest practical receivers consists of a tuned circuit for selecting the desired radio signal and a detector for separating the modulation from the carrier. The detector may be either a mineral such as galena or carborundum, or else a vacuum tube. Figure 1 shows such a receiver using a diode vacuum tube as a detector. The *sensitivity* of this receiver, or in other words its ability to make audible weak signals, would be very low, but it is useful to illustrate the basic action of all receivers.

Energy from a passing radio wave will induce a voltage in the antenna and cause a radio-frequency current to flow from antenna to ground through coil L_1. The alternating magnetic field set up around L_1 links with the turns of L_2 and causes an r.f. current to flow through the parallel-tuned circuit. L_2-C. When variable condenser C is adjusted so that the tuned circuit is resonant at the frequency of the applied signal, the r.f. voltage is maximum, as explained in Chapter 2. This r.f. voltage is applied to the diode detector where it is rectified into a pulsating direct current and passed through the earphones. The pulsations in this voltage correspond to the modulation placed on the signal at the transmitter. As the earphone diaphragms vibrate back and forth following the pulsating current they audibly reproduce the original modulation.

The operation of the detector circuit is shown graphically above the detector circuit in Figure 1. The modulated carrier is shown at A, as it is applied to the antenna. B represents the same carrier, increased in amplitude, as it appears across the tuned circuit. In C the pulsating d.c. output from the detector is seen.

The receiver shown in Figure 1 would be an extremely poor one, being suitable only for use in the immediate vicinity of a transmitting station. By adding an audio amplifier, however, as shown in Figure 2A, the output of the receiver may be increased greatly. In 2A, the earphones of Figure 1 have been replaced by a resistor, R, and an r.f. by-pass condenser, C_1. The audio voltage across R and C_1 is coupled to the grid of a class A audio amplifier by means of a coupling condenser C_2, and the headphones are placed in the plate circuit of the amplifier stage. Grid bias is supplied by a C battery, which is connected to the amplifier grid through a high resistance, R_1.

To simplify the circuit shown at 2A, the load resistor, R, and its by-pass condenser may be moved around the circuit until they are in series with the diode plate, instead of its cathode. The voltage across R and C_1 is still pulsating d.c., with the pulsation corresponding to the modulation on the signal, but the d.c. voltage at the diode plate is now always negative in respect to ground. Having a negative voltage at the diode plate allows the amplifier stage grid to be directly connected to this point, thus dispensing with the bias battery, the grid return resistor R_1, and coupling condenser C_1.

Still further simplification of the circuit is shown at 2C, where the triode grid has entirely replaced the diode plate, thus eliminating one tube from the circuit. An r.f. by-pass condenser C_3 has been added in 2C to remove any r.f. which finds its way into the plate circuit. The circuit shown at 2C is known as a *grid leak detector*, and as the above discussion has shown, it is simply a diode detector plus an audio amplifier, both combined in a single tube. The grid-leak detector is not limited to tri-

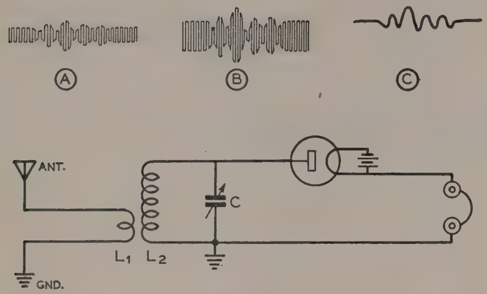


Figure 1.
DIODE DETECTOR RECEIVER.
This circuit would make a very poor receiver, but it is useful in illustrating the process of detection. The operation of the circuit is described in the text.

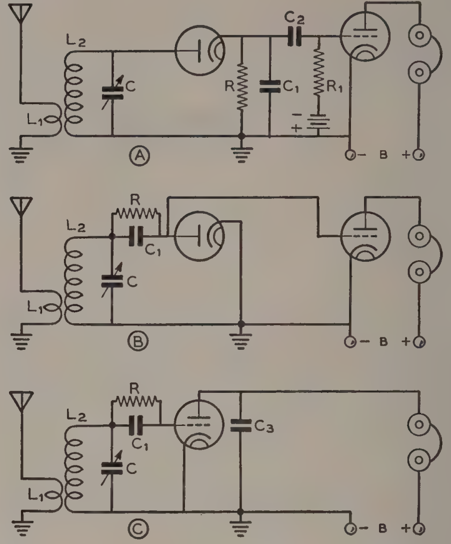


Figure 2.
GRID-LEAK DETECTOR DEVELOPMENT.
Showing the development of a grid-leak detector from a diode detector plus audio amplifier.

odes; tetrodes or pentodes may also be used, these generally having greater sensitivity than the triodes.

Since the grid-leak detector shown in Figure 2C produces an audio output that corresponds only to the modulation on a signal, it is useless for the reception of unmodulated radiotelegraph (c.w.) signals. Such signals may be made audible, however, by applying another radio-frequency signal to the detector along with the desired signal. If the difference in frequency between the two signals falls in the audio-frequency range, a heterodyne or "beat note" at the difference frequency will be heard in the 'phones. The beat note will, of course, start and stop to conform with the code characters of the distant signal.

The local signal which is used to beat with the desired c.w. signal in the detector may be supplied by a separate low-power oscillator in the receiver itself, or the detector may be made to self-oscillate, and thus serve the dual purpose of detector and oscillator. A detector which self-oscillates to provide a beat note is known as an *autodyne* detector, and the process of obtaining feedback between the detector plate and grid is called *regeneration*. A typical regenerative detector is shown in Figure 3.

In the circuit shown in Figure 3, radio-frequency energy from the detector plate is coupled back to the grid by means of the tickler coil, L_2. This coil is closely coupled to the grid winding, L_1, and when the phasing between the two coils is correct, the energy from the plate circuit adds to that in the grid circuit, thus increasing the amplification. When the energy fed back is sufficient to overcome the losses in the grid circuit, oscillation takes place and autodyne reception occurs.

The autodyne detector is most sensitive when it is barely oscillating, and for this reason a regeneration control is always included in the circuit to adjust the feedback to the proper amount. Condenser C_2 in Figure 3 is the regeneration control. This condenser serves as a variable plate by-pass condenser, and it is commonly called a "throttle condenser."

With the detector regenerative, that is, with feedback taking place, but not enough to cause oscillation, it is also extremely sensitive. When the circuit is adjusted to operate in this manner, modulated signals may be received with considerably greater strength than when the detector is in a non-regenerative condition.

Other Regenerative Detectors. The circuit shown in Figure 3 is by no means the only one which will give satisfactory results as a regenerative detector. There are several methods by which regeneration may be obtained, and also several alternative methods of controlling the regeneration. In tubes with an indirectly-heated cathode, regeneration may be obtained by tapping the cathode onto the grid coil a few turns up from the ground end, or by returning the cathode to ground through a coil coupled to the grid winding. With tetrode or pentode tubes, feedback is sometimes provided by connecting the screen, rather than the plate, to the tickler coil.

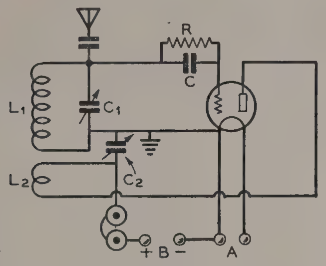


Figure 3.
TRIODE REGENERATIVE DETECTOR.
The regenerative detector makes the simplest practical high-frequency receiver.

Alternative methods of controlling regeneration consist of providing means for varying the voltage on one of the tube elements, usually the plate or screen. Examples of some of the possible variations in regeneration and control methods are shown in Figure 4.

Amplifier Stages

The sensitivity and selectivity of the receiver may be increased by adding a tuned radio-frequency amplifier between the detector and the antenna. The radio-frequency (r.f.) amplifier stage increases the strength of the signal applied to the detector, and thus the receiver with an r.f. stage is capable of giving a useful audio output on signals much weaker than those which represent the minimum useful level of signal strength for the detector alone. The addition of the tuned circuits required in the r.f. amplifier also improves the selectivity of the receiver.

Audio frequency amplifiers may be added after the detector to enable weak signals which have been detected to be amplified sufficiently to actuate the sound-producing mechanism in the headphones or speaker.

Radio Frequency Amplifiers. A typical tuned radio-frequency amplifier connected ahead of a regenerative detector is shown in Figure 5. A pentode tube is used in the r.f. stage with a tuned grid circuit and inductive coupling from the antenna and to the detector. Capacitive coupling could be used in both instances; but in the case of the coupling between stages, a high-impedance radio-frequency choke would have to be connected to the plate of the r.f. stage to allow plate voltage to be applied to the tube. A capacity-coupling system which allows the r.f. choke to be dispensed with is shown in Figure 6. This circuit is often used at ultra-high frequencies where a high-impedance resonant circuit in the plate of the r.f. tube is desired in order to obtain greater amplification.

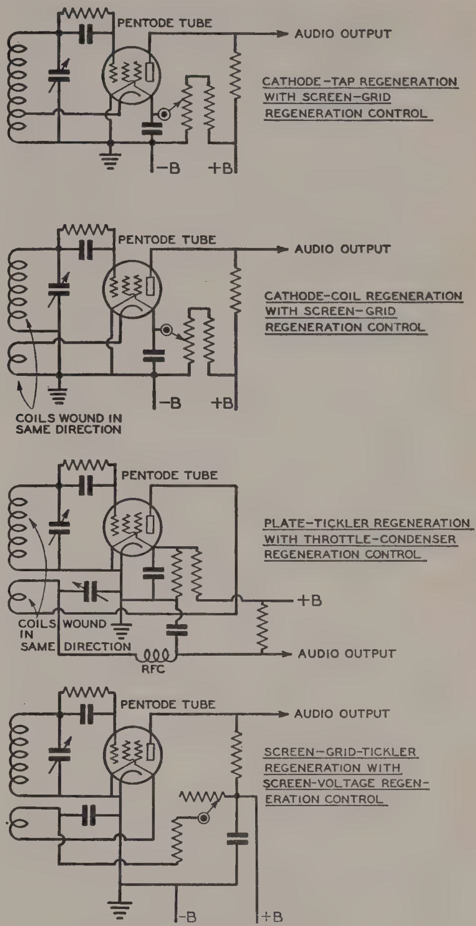


Figure 4.
REGENERATIVE DETECTOR CIRCUITS.
These circuits illustrate some of the more popular regenerative detectors. Values of 1 to 3 megohms for grid leaks are common. The grid condenser usually has a capacity of .0001 μfd., while the screen by-pass is 0.1 μfd. Pentode detectors operate best when the feedback is adjusted so that they start to oscillate with from 30 to 50 volts on the screen grid.

The dotted line running between condensers C_1 and C_2 in Figure 5 indicates that their rotor shafts are mechanically connected (or *ganged*) together so that both tuned circuits may be resonated to the desired signal with but a single dial. When the r.f. stage is separate from the receiver, and its tuning control is not ganged with that of the receiver proper, it is commonly known as a *preselector*.

A preselector may be added to any receiver, but it is most often used with the superheterodyne type.

The amplification obtained in an r.f. stage depends upon the type of circuit which is used; if the plate load impedance can be made very high, the gain may be as much as 200 or 300 times. Normal values of gain in the broadcast band are in the vicinity of 50 times. A gain of 30 per r.f. stage is considered excellent for shortwave receivers which have a range of from 30 to 100 meters. Radio-frequency amplifiers for the very short wavelengths, such as from 5 to 20 meters, seldom provide a gain of more than 10 times, because of the difficulty in obtaining high load impedances and the shunt effect of the rather high input capacities of most ordinary screen-grid tubes.

Regenerative R.F. Stages. In low cost receivers, and in those where maximum performance with a minimum number of stages is desired, controlled regeneration in an r.f. stage is often used. The regenerative r.f. amplifier increases amplification and selectivity in a manner similar to that of the regenerative detector. The regenerative r.f. amplifier is never allowed to oscillate, however; the greatest amplification is obtained with the circuit operating just below the point of oscillation. Figure 7 shows a regenerative r.f. stage of the type generally used on the higher frequencies.

One minor disadvantage of the regenerative r.f. stage is the need for an additional control for regeneration. A more important disadvantage is that, due to the high degree of selectivity obtainable with the regenerative stage, it is usually impossible to secure accurate enough tracking between its tuning circuit and the other tuning circuits in the receiver to make single-dial control feasible. Where single-dial control is desired, a small "trimmer" condenser is usually provided across the main r.f.-stage tuning condenser. By making this condenser operable from the front panel, it is possible to compensate manually for slight inaccuracies in the tracking. A further discussion of regenerative r.f. stages will be found under the section on superheterodyne receivers, in which connection they are most often used.

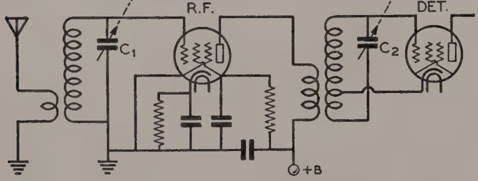


Figure 5.
R.F. AMPLIFIER CIRCUIT.
An r.f. amplifier ahead of the regenerative detector will increase the receiver's selectivity and sensitivity.

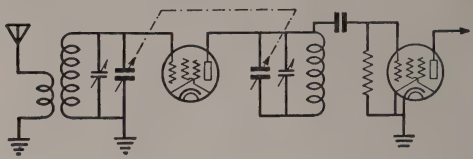


Figure 6.
CAPACITY COUPLING BETWEEN STAGES.
This type of coupling circuit is often used at ultra-high frequencies when it is desired to have a high impedance plate load for the r.f. stage.

Audio Amplifiers. Audio amplifiers are employed in nearly all radio receivers. The audio amplifier stage or stages are usually of the class A type, although small class B stages are used in some receivers. The operation of both of these types of amplifiers was described in Chapter 3. The purpose of the audio amplifier is to bring the relatively weak signal from the detector up to a strength sufficient to operate a pair of headphones or a loud speaker. Either triodes, pentodes, or beam tetrodes may be used, the pentodes and beam tetrodes usually giving greater output. In some receivers it is possible to operate the headphones directly from the detector, without audio amplification. In such receivers, a single audio stage with a beam tetrode or pentode tube is ordinarily used to drive the loud speaker. Representative audio amplifier arrangements will be found in Chapter 6.

Superregenerative Receivers

At ultra-high frequencies, when it is desired to keep weight and cost at a minimum, a special form of the regenerative receiver known as the *superregenerator* is often used. The superregenerator is essentially a regenerative receiver with a means provided to throw the detector rapidly in and out of oscillation. The frequency at which the detector is made to go in and out of oscillation varies in different receivers, but is usually between 20,000 and 100,000 times a second. As a consequence of the sensitivity of the oscillating detector being increased, the usual "regeneration hiss" is greatly amplified when no signal is being received. This hiss diminishes in proportion to the strength of the received signal, loud signals eliminating the hiss entirely.

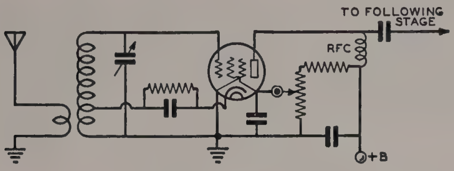


Figure 7.
REGENERATIVE R.F. AMPLIFIER.
The use of regeneration in the r.f. amplifier allows greater amplification to be obtained, but the increased gain is accompanied by an increase in tube noise.

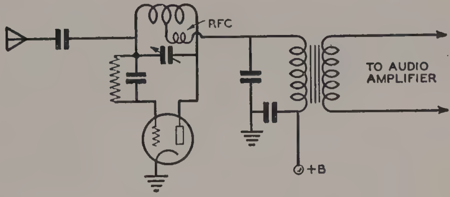


Figure 8.
SUPERREGENERATIVE DETECTOR.
This extremely sensitive self-quenched detector arrangement is often used at ultra-high frequencies. The plate blocking condenser must have low reactance at the quench frequency; a value of .006 μfd. is common.

Detector Operation. There are two systems in common use for causing the detector to break in and out of oscillation rapidly. In one, a separate *interruption-frequency* oscillator is arranged so as to vary the voltage rapidly on one of the detector tube elements (usually the plate, sometimes the screen) at the high rate necessary. The interruption-frequency oscillator commonly uses a conventional tickler-feedback circuit with coils appropriate for the frequency at which it operates.

The second, and simplest, type of superregenerative detector circuit is arranged so as to produce its own interruption frequency oscillation, without the aid of a separate tube. The detector tube damps (or "quenches") itself out of signal-frequency oscillation at a high rate by virtue of the use of a high value of grid leak and proper size plate-blocking and grid condensers. In this type of "self-quenched" detector, the grid leak is quite often returned to the positive side of the power supply (through the coil) rather than to the cathode. A representative self-quenched superregenerative detector circuit is shown in Figure 8.

Both types of superregenerative detectors act as small transmitters and radiate broad, rough signals unless they are well shielded and preceded by an r.f. stage. For this reason they are not too highly recommended for use on frequencies below 60 Mc. However, there are occasionally cases where their use is justified on the 56-to-60 Mc. band. The superregenerative receiver tunes very broadly, receiving a band at least 100 kc. wide. For this reason it is widely popular for the reception of unstable, modulated oscillators at ultra-high frequencies.

Frequency modulation reception is possible with superregenerative receivers, although with the amount of "swing" ordinarily used in frequency-modulated transmitters, the audio output of the receiver is comparable to that obtained when the signal is amplitude modulated at a rather low percentage. If a relatively wide swing is used in the transmitter, however, the audio output of the receiver will compare favorably with that obtained from a fully amplitude modulated carrier of equivalent strength.

Practical superregenerative receiver circuits, along with a further discussion of their operation, will be found in Chapter 18.

Superheterodyne Receivers

Because of its superiority and nearly universal use in all fields of radio reception except at the extremely high "micro wave" frequencies, the theory of operation of the superheterodyne should be familiar to every radio experimenter, whether or not he contemplates building a receiver of this type. The following discussion concerns superheterodynes for amplitude-modulation reception. It is, however, applicable in part to receivers for frequency modulation. The points of difference between the two types of receivers, together with circuits required for f.m. reception, will be found in Chapter 9.

Principle of Operation. In the superheterodyne, a radio-frequency circuit is tuned to the frequency of the incoming signal, and the signal across this circuit applied to a vacuum-tube mixer stage. In the mixer stage, the signal is mixed with a steady signal generated in the receiver, with the result that a signal bearing all the modulation applied to the original *but of a frequency equal to the difference between the local oscillator and incoming signal frequencies* appears in the mixer output circuit. The output from the mixer stage is fed into a fixed-tune *intermediate-frequency amplifier*, where it is amplified and detected in the usual manner, and passed on to the audio amplifier. Figure 9 shows a block diagram of the fundamental superheterodyne arrangement.

Superheterodyne Advantages. The advantages of superheterodyne reception are directly attributable to the use of the fixed-tune intermediate-frequency (i.f.) amplifier. Since all signals are converted to the intermediate frequency, this section of the receiver may be designed for optimum selectivity and amplification without going into the extremely complicated tunable band pass arrangements or the number of stages which would be necessary if the signal-frequency tuning circuits were designed to have a comparable degree of selectivity and gain. High amplification is easily obtained in the intermediate-frequency amplifier, since it operates at a relatively low frequency, where conventional pentode-type tubes give a great deal of voltage gain. A typical i.f. amplifier stage is shown in Figure 10.

From the diagram it may be seen that both the grid and plate circuits are tuned. Tuning both circuits in this way is advantageous in two ways; it increases the selectivity, and it allows the tubes to work into a high-impedance resonant plate load, a very desirable condition where high gain is desired. The tuned circuits used for coupling between i.f. stages are known as *i.f. transformers.* These will be more fully discussed later in this chapter.

Choice of Intermediate Frequency. The choice of a frequency for the i.f. amplifier involves several considerations. One of these considerations is in the matter of selectivity; as a general rule, the lower the intermediate frequency the better the selectivity. On the other hand, a rather high intermediate frequency is desirable from the standpoint of *image* elimination, and also for the reception of signals from television and f.m. transmitters and modulated self-controlled oscillators, all of which occupy a rather wide band of frequencies, making a broad selectivity characteristic desirable. Images are a peculiarity common to all superheterodyne receivers, and for this reason they are given a detailed discussion later in this chapter.

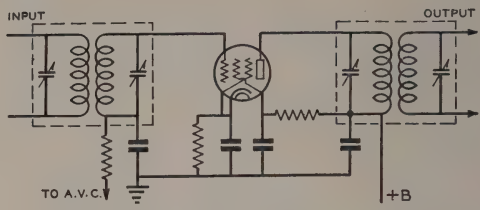


Figure 10.
TYPICAL I.F. AMPLIFIER STAGE.
Variable-μ pentodes are ordinarily used as i.f. amplifier tubes. Most of the ordinary tubes require a cathode resistor of around 300 ohms and a 100,000-ohm screen dropping resistor. The high-transconductance "television" type pentodes usually need less cathode resistance, and values as low as 100 ohms are common. The screen resistor for the "television" types may have a value between 50,000 and 75,000 ohms. By-pass condensers are usually .05 or 0.1-μfd.

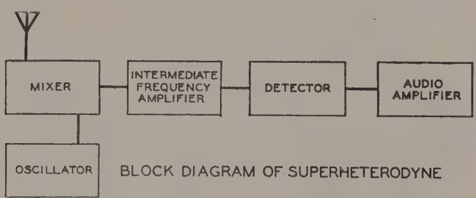


Figure 9.
THE ESSENTIAL PARTS OF A SUPERHETERODYNE RECEIVER.
There are several possible variations of this arrangement. R.f. amplifier stages often are used ahead of the mixer. Occasionally the i.f. amplifier stages are omitted in simple superheterodynes.

While intermediate frequencies as low as 30 kc. were common at one time, and frequencies as high as 20,000 kc. are used in some specialized forms of receivers, most present-day communications superheterodynes nearly always use intermediate frequencies around either 455 kc. or 1600 kc. Two other frequencies which are sometimes encountered in broadcast-band receivers are 175 kc. and 262 kc.

Generally speaking, it may be said that for maximum selectivity consistent with a reasonable amount of image rejection for signal frequencies up to 30 Mc., intermediate frequencies in the 450–470 kc. range are used, while for a good compromise between image rejection and selectivity the i.f. amplifier will often operate at 1600 kc. For the reception of both amplitude and frequency modulated signals above 30 Mc., intermediate frequencies near 2100, 4300 and 5000 kc. are most often used. The intermediate amplifiers in television receivers will usually be found to operate in the region between 8000 and 15,000 kc.

Arithmetical Selectivity. Aside from allowing the use of fixed-tune band pass amplifier stages, the superheterodyne has an overwhelming advantage over the t.r.f. type of receiver because of what is commonly known as *arithmetical selectivity.*

This can best be illustrated by considering two receivers, one of the t.r.f. type and one of the superheterodyne type, both attempting to receive a desired signal at 10,000 kc. and eliminate a strong interfering signal at 10,010

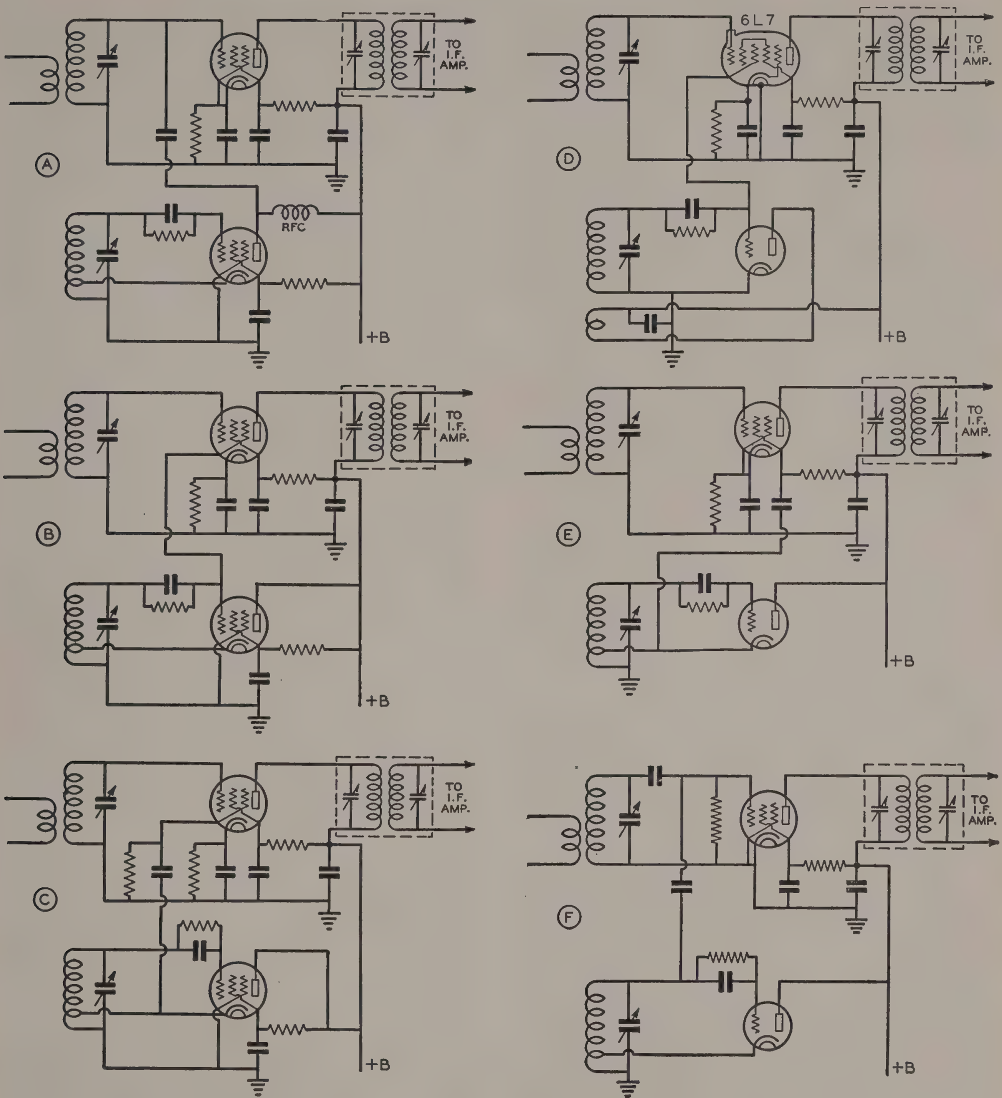


Figure 11.
MIXER-OSCILLATOR COMBINATIONS.
The various oscillators do not have to be used with the mixers with which they happen to be shown. The triode oscillator shown at E could replace the pentode circuit shown at B, for instance.

kc. In the t.r.f. receiver, separating these two signals in the tuning circuits is practically impossible, since they differ in frequency by only 0.1 per cent. However, in a superheterodyne with an intermediate frequency of, for example, 1000 kc., the desired signal will be converted to a frequency of 1000 kc. and the interfering signal will be converted to a frequency of 1010 kc., both signals appearing at the input of the i.f. amplifier. In this case, the two signals may be separated much more readily, since they differ by 1 per cent, or 10 times as much as in the first case.

Mixer Circuits. The most important single section of the superheterodyne is the *mixer*. No matter how much signal is applied to the mixer, if the signal is not converted to the intermediate frequency and passed on to the

i.f. amplifier with a strength greater than the noise level at the i.f. input, it is lost. The tube manufacturers have released a large variety of special tubes for mixer applications, and these, as well as improved circuits with older type tubes, have resulted in highly effective mixer arrangements in present-day receivers.

Figure 11 shows several representative mixer-oscillator circuits. At "A" is illustrated control-grid *injection* from an electron-coupled oscillator to the mixer. The mixer tube for this type of circuit is usually a remote-cut-off pentode of the 57—6J7 type. The coupling condenser, C, between the oscillator and mixer is quite small, usually 1 or 2 $\mu\mu$fd. Quite often sufficient coupling will be obtained by simply twisting a turn or two of hook-up wire around the mixer grid lead, and using the capacity between the two wires as a coupling condenser.

This same circuit may be used with the oscillator output being taken from the oscillator grid or cathode. The only disadvantage to this method is that interlocking, or "pulling," between the mixer and oscillator tuning controls is likely to take place. A rather high value of cathode resistor (10,000 to 50,000 ohms) is usually used with this circuit.

Injection of oscillator voltage into mixer elements other than the control grid, is illustrated by Figures 11B, C, D and E. The circuit of 11B shows injection into the suppressor grid of the mixer tube. The suppressor is biased negatively by connecting it directly to the grid of the oscillator.

An alternative method of obtaining bias for the suppressor, and one which is less prone to cause interlocking between the oscillator and mixer, is shown in Figure 11C. In this circuit, the suppressor bias is obtained by allowing the rectified suppressor-grid current to flow through a 50,000- or 100,000-ohm resistor to ground. The coupling condenser between oscillator and mixer may be 50 or 100 $\mu\mu$fd. with this circuit, depending upon the frequency. Output from the oscillator may be taken from the cathode instead of the grid end of the coil, as shown, if sufficient oscillator output is available. Mixer cathode resistors having values between 500 and 5000 ohms are ordinarily used with the circuits of 11B and C.

The mixer circuit shown in 11D is similar in appearance to that of 11B. The difference in the two lies in the type of tube used as a mixer. The 6L7 shown in 11D is especially designed for mixer service. It has a separate shielded *injector grid,* by means of which voltage from the oscillator may be injected. This circuit permits the same variations as the suppressor-injection system in regard to the method of connection into the oscillator circuit. The 6L7 requires rather high screen voltage and draws considerable screen current, and, for these reasons, the screen-dropping resistor is usually made around 10,000 or 15,000 ohms, which is considerably less than the values of 50,000 to 100,000 ohms used with most other mixer tubes.

Figure 11E shows injection into the mixer screen grid. When connected in the manner shown, a rather large (.01 to 0.1 μfd.) coupling condenser may be used. This circuit is likely to cause rather bad pulling at high frequencies, as there is no electrostatic shielding within the mixer tube between the screen grid and the control grid. A variation of this circuit, in which the pulling effect is reduced considerably, consists of using an electron-coupled oscillator circuit similar to that shown in 11A and connecting the plate of the oscillator and the screen of the mixer directly together. A voltage of about 100 volts is then applied to both the oscillator plate and the mixer screen.

E.C.O. Harmonics. One disadvantage to the use of an electron-coupled type oscillator with the output taken from the plate which should be borne in mind by the constructor, is that the untuned plate circuit of the e.c. oscillator contains a large amount of harmonic output. Therefore, considerable selectivity must be used ahead of the mixer to prevent the harmonics of the oscillator from beating with undesired signals at higher frequencies and bringing them in along with the desired signal. If it is desired to use an e.c. type oscillator to secure receiver stabilization in regard to voltage changes, it will usually be found best to take the oscillator output from the tuned grid circuit, where the harmonic content is low. The plate of the oscillator tube may be by-passed directly to ground when this arrangement is used.

Improved Control-Grid Injection. In Figure 11F an improved control-grid injection type mixer circuit is shown. This circuit allows peak mixer conversion transconductance under wide variations in oscillator output. The bias on the mixer is automatically maintained at the correct value through the use of grid-leak bias, rather than by the more common cathode bias arrangement. The mixer grid leak should have a value of from 3 to 5 megohms. As in the circuit shown at 11A, the coupling condenser should be quite small—on the order of 1 or 2 $\mu\mu$fd. It is absolutely essential that a rather high value of series screen dropping resistor be used with this circuit to limit the current drawn by the mixer tube in case the oscillator injection voltage, and consequently the mixer bias, is inadvertently removed. The value of the screen resistor will

probably lie around 100,000 ohms or above, depending upon the type of mixer tube and the available plate voltage. The resistor value should be determined experimentally by using a value which limits the mixer cathode current when the oscillator is not operating to the maximum permissible current specified by the tube manufacturer.

The different oscillator circuits shown in Figure 11 are not necessarily limited to use with the mixers with which they happen to be shown. Almost any oscillator arrangement may be used with a particular mixer circuit. Examples of some of the possible combinations will be found in Chapter 6.

Converter Tubes. There is a series of *pentagrid converter* tubes available in which the functions of the oscillator and mixer are combined in a single tube. Typical of these tubes are the 6A7, 6A8, and 6SA7. The term *pentagrid* has been applied to these tubes because they have 5 grids, one of the extra grids being used as grid and the other as the anode for the oscillator section of the circuit. Suitable circuits for use with these tubes are shown in Figures 12A and 12B.

Dual Unit Converters. Another set of combination tubes known as *triode-heptodes* and *triode-hexodes* is also available for use as combination mixers and oscillators. These tubes are exemplified by the 6J8G and the 6K8; they get their name from the fact that they contain two separate sets of elements—a triode and a heptode in one case, and a triode and a hexode in the other. Representative circuits for both types of tube are shown at 12C and 12D.

Separate Oscillator. Certain of the combination mixer-oscillator tubes make exceptionally good high frequency mixers when their oscillator section is left unused and the oscillator section grid is connected to a separate oscillator capable of high output. The 6K8, 6J8G and 6SA7 perform particularly well when used in this manner. A circuit of this type for use with a 6K8 is shown in Figure 13. The points marked "X" in Figure 12 show the proper place to inject r.f. from a separate oscillator with the other combination type converter tubes. When the 6A7 and 6A8 types are used with a separate oscillator, the unused oscillator anode-grid is connected directly to the screen.

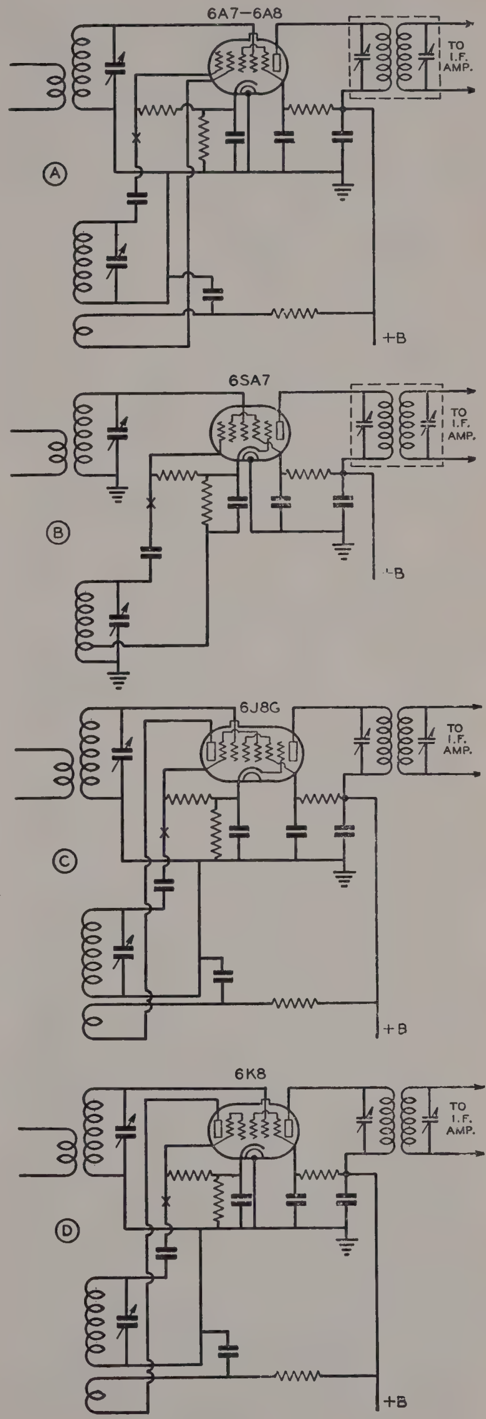


Figure 12.
CONVERTER CIRCUITS.
A and B are for "pentagrid" tubes, and C and D are for "triode-heptode" and "triode-hexode" tubes. The points marked "X" show where injection from a separate oscillator may be introduced.

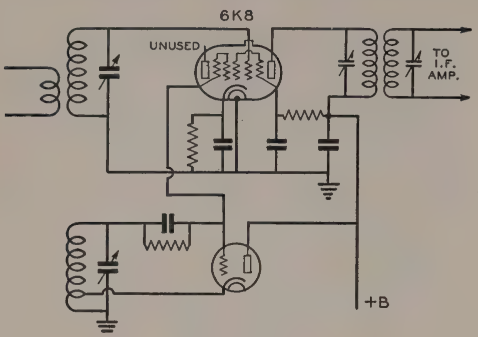


Figure 13.
USING A SEPARATE OSCILLATOR WITH A DUAL-PURPOSE CONVERTER TUBE.
Certain of the better dual-purpose converter tubes make excellent mixers when used with a separate oscillator. The points marked "X" in Figure 12 show where a separate oscillator may be connected with each of the tubes shown.

Mixer Noise and Images

The effects of *mixer noise* and *images* are troubles common to all superheterodynes, and since both these effects can largely be obviated by the same remedy, they will be considered together.

Mixer Noise. Mixer noise of the shot-effect type, which is evidenced by a hiss in the audio output of the receiver, is caused by exceedingly small irregularities in the plate current in the mixer stage. Noise of an identical nature is generated in the amplifier stages of the receiver, but due to a certain extent to the fact that the gain in the mixer stage is considerably lower than in an amplifier stage using the same tube, the proportion of inherent noise present in a mixer usually is considerably greater than in an amplifier stage.

Although this noise cannot be eliminated, its effects can be greatly minimized by placing sufficient signal-frequency amplification having a high signal-to-noise ratio ahead of the mixer. This remedy causes the signal output from the mixer to be large in proportion to the noise. Increasing the gain after the mixer will be of little advantage in eliminating mixer noise difficulties; greater selectivity after the mixer will help to a certain extent, but cannot be carried too far, since this type of selectivity decreases the i.f. bandpass and reduces the strength of the high-frequency components of modulated signals.

Images. Images are a result of frequency conversion. They are a consequence of the fact that there are two signal frequencies which will combine with a single oscillator frequency to produce the same difference frequency. For example: a superheterodyne with its oscillator operating on a higher frequency than the signal, which is common practice in present superheterodynes, is tuned to receive a signal at 14,100 kc. Assuming an i.f.-amplifier frequency of 450 kc., the mixer input circuit will be tuned to 14,100 kc., and the oscillator to 14,100 plus 450, or 14,550 kc. Now, a strong signal at the oscillator plus the intermediate frequency (14,550 plus 450, or 15,000 kc.) will also give a difference frequency of 450 kc. in the mixer output and will be received just as though it were actually on 14,100 kc., the frequency of the desired signal. The image is always *twice* the intermediate frequency away from the desired signal.

The only way that the image could be eliminated in this particular case would be to make the selectivity of the mixer input circuit, and any circuits preceding it, great enough so that the 15,000-kc. signal would be eliminated with these circuits tuned to 14,100 kc.

For any particular intermediate frequency, image interference troubles become increasingly greater as the frequency to which the signal-frequency portion of the receiver is tuned is increased. This is due to the fact that the percentage difference between the desired frequency and the image frequency decreases as the receiver is tuned to a higher frequency. The ratio of strength between a signal at the image frequency and a signal at the frequency to which the receiver is tuned required to give equal output, is known as the *image ratio*. The higher this ratio, the better the receiver in regard to image-interference troubles.

With but a single tuned circuit between the mixer grid and the antenna, and with 400–500 kc. i.f. amplifiers, image ratios of one hundred and over are easily obtainable up to frequencies around 5000 kc. Above this frequency, greater selectivity in the mixer grid circuit (through the use of regeneration) or additional tuned circuits between the mixer and the antenna, are necessary if a good image ratio is to be maintained.

R.F. Stages. Since the necessary tuned circuits between the mixer and the antenna can be combined with tubes to form r.f. amplifier stages, the reduction of the effects of mixer noise and the increasing of the image ratio can be accomplished in a single section of the receiver. When incorporated in the receiver, this section is known simply as an *r.f. amplifier;* when it is a separate unit with a separate tuning control it is often known as a *preselector*. Either one or two stages are commonly used in the preselector or r.f. ampli-

fier. Some single-stage preselectors and a few 2-stage units use regeneration to obtain still greater amplification and selectivity.

Double Conversion. As previously mentioned, the use of a higher intermediate frequency will also improve the image ratio, at the expense of i.f. selectivity, by placing the desired signal and the image farther apart. To give both good image ratio at the higher frequencies and good selectivity in the i.f. amplifier, a system known as *double conversion* is sometimes employed. In this system, the incoming signal is first converted to a rather high intermediate frequency, such as 1600 kc., and then amplified and again converted, this time to a much lower frequency, such as 175 kc. The first i.f. frequency supplies the necessary wide separation between the image and the desired signal, while the second one supplies the bulk of the i.f. selectivity.

Regenerative Preselectors. R.f. amplifiers for wave-lengths down to 20 meters can be made to operate efficiently in a nonregenerative condition. The amplification and selectivity are ample over this range. For higher frequencies, on the other hand (wave-lengths below 20 meters), *controlled regeneration* in the r.f. amplifier is often desirable for the purpose of increasing the gain and selectivity.

As previously discussed, a disadvantage of the regenerative r.f. amplifier is the need for an additional regeneration control, and the difficulty of maintaining alignment between this circuit and the following tuned circuits. Resonant effects of antenna systems usually must be taken into account; a variable antenna coupling device can sometimes be used to compensate for this effect, however.

The reason for using regeneration at the higher frequencies and not at the medium and low frequencies can be explained as follows: The signal-to-noise ratio (output signal) of the average r.f. amplifier is not made higher by the incorporation of regeneration, but the signal-to-noise ratio of the *receiver as a whole* is improved at the very high frequencies because of the extra gain provided ahead of the mixer, this extra gain tending to make the signal output a larger portion of the total signal-plus-noise output of the receiver. At low frequencies an r.f. stage has sufficient gain to do this without resorting to regeneration.

Signal-Frequency Tuned Circuits

The signal-frequency tuned circuits in superheterodynes and tuned radio frequency types of receivers consist of coils of either the solenoid or universal-wound types shunted by variable condensers. It is in these tuned circuits that the causes of success or failure of a receiver often lie. The universal-wound type coils usually are used at frequencies below 2000 kc.; above this frequency the single-layer solenoid type of coil is more satisfactory.

Impedance and Q. The two factors most affecting the tuned circuits are impedance and Q. As explained in Chapter 2, Q is the ratio of reactance to resistance in the circuit. Since the resistance of modern condensers is low at ordinary frequencies, the resistance usually can be considered to be concentrated in the coil. The resistance to be considered in making Q determinations is the r.f. resistance, not the d.c. resistance of the wire in the coil. The latter ordinarily is low enough that it may be neglected. This r.f. resistance is influenced by such factors as wire size and type, and the proximity of metallic objects or poor insulators, such as coil forms with high losses. It may be seen from the curves shown in Chapter 2 that higher values of Q lead to better selectivity and increased r.f. voltage across the tuned circuit. The increase in voltage is due to an increase in the circuit impedance with the higher values of Q.

Frequently it is possible to secure an increase in impedance in a resonant circuit, and consequently an increase in gain from an amplifier stage, by increasing the reactance through the use of larger coils and smaller tuning condensers (higher L/C ratio). In spite of the fact that the Q of the coil probably will be lowered by this process, the impedance, which is a function of both reactance and Q, will be greater because for small increases in reactances the reactance will increase faster than the Q decreases. The selectivity will be poorer, but in superheterodyne receivers selectivity in the signal-frequency circuits is of minor importance where signals on adjacent channels are concerned. On the other hand, the t.r.f. type of receiver requires good selectivity in the tuned circuits, and a compromise between impedance and Q must be made.

Input Resistance. Another factor which influences the operation of tuned circuits is the input resistance of the tubes placed across these circuits. At broadcast frequencies, the input resistance of most conventional r.f. amplifier tubes is high enough so that it is not bothersome. But as the frequency is increased, the input resistance becomes lower and lower, until it ultimately reaches a value so low that no amplification can be obtained from the r.f. stage. The two contributing factors to the decrease in input resistance with increasing frequency are the transit time required by an electron traveling between the cathode and grid, and the inductance of the cathode lead common to both the plate and grid circuits. As the frequency becomes higher, the transit time can become an appreciable portion of the

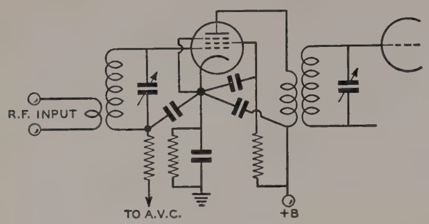


Figure 14.
BY-PASSING IN HIGH-FREQUENCY STAGES.

To reduce the detrimental effects of common cathode inductance at the higher frequencies, all by-pass condensers should be returned directly to the socket cathode terminal. Certain of the newer type tubes have two cathode terminals, and with these types the plate and screen by-passes should be returned to one terminal and the grid by-pass to the other.

time required by an r.f. cycle of the signal voltage, and current will actually flow into the grid even though it is biased negatively. The result of this effect is similar to that which would be obtained by placing a resistance between the tube's grid and cathode.

Since the input resistance of conventional tubes can reach rather low values at frequencies above 10 Mc. or thereabouts, there is often no practical advantage to be realized by going to great pains to design a very high impedance tuned circuit for these frequencies, and then shunting it with the tube's input resistance. At any given frequency the tube input resistance remains constant, regardless of what is done to the tuned circuit, and increasing the tuned circuit impedance beyond three or four times the input resistance will have but little effect on the net grid-to-ground impedance of the amplifier stage.

The limiting factor in r.f. stage gain is the ratio of input conductance to the tube transconductance. When the input conductance becomes so great that it equals the transconductance, the tube no longer can act as an amplifier. One of the ways of increasing the ratio of transconductance to input conductance is exemplified by the "acorn" and "miniature" type tubes, in which the input conductance is reduced through the use of a smaller element structure while the transconductance remains nearly the same as that of tubes ordinarily used at lower frequencies. Another method of accomplishing an increase in transconductance-input conductance ratio is by greatly increasing the transconductance at the expense of a proportionately small increase in input conductance. The latter method is exemplified by the so-called "television pentodes," which have extremely high transconductance and an input conductance several times that of the acorn tubes.

A recently-released tube, which will probably be followed by others of the same general type, gives an increase in transconductance-input conductance ratio by the use of separate cathode leads for the grid and plate returns in conjunction with a design which gives fairly high transconductance. By using separate leads to the cathode for the input-output circuit return connections, the inductance common to both circuits may be held to a minimum, and the input conductance thus decreased.

With conventional tubes having a single cathode terminal, the only control the constructor has over the input resistance is through eliminating, so far as possible, the cathode lead inductance common to the input and output circuits. This means that all by-pass condensers associated with a tube should be connected separately and directly to the socket cathode terminal. The ground connection for the stage may be made by a single condenser from the cathode to chassis. A typical circuit is shown in Figure 14.

Some of the difficulties presented by input-resistance effects may be obviated by tapping the grid down on the coil, as shown in Figure 15. Although this circuit does not actually cause any reduction in the tube's input conductance, it does remove some of the loading from the tuned circuit, and thus will improve the selectivity. With a tuned circuit, which by itself has a high impedance, there will be no loss in r.f. voltage applied to the grid, and the net result of tapping the grid down on the coil will be an improvement in selectivity without a loss in stage gain. This circuit is commonly employed with high-transconductance tubes when operating on the 28–30 Mc. amateur band, and nearly always with such tubes on the 56–60 Mc. band. Acorn and u.h.f. "miniature" tubes, due to their smaller dimensions and lower capacities, are considerably better than the conventional types at ultra-high frequencies, and it usually will not be found necessary to tap their grids down on the tuned circuit until frequencies around 200 Mc. are reached.

Superheterodyne Tracking. Because the mixer (and r.f. stages, if any) and the oscillator operate on different frequencies in superheterodynes, in some cases it is necessary to make special provisions to allow the oscillator to track with the other tuned circuits when similar tuning condensers are used. The usual method of obtaining good tracking is to operate the oscillator on the high-frequency side of the mixer and use a series "tracking condenser" to slow down the tuning rate of the

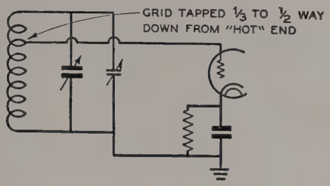


Figure 15.
REDUCING GRID-LOADING EFFECTS.
By tapping the grid down on the coil, as shown, the selectivity may be increased when high-transconductance tubes are used at high frequencies.

oscillator. The oscillator tuning rate must be slower because it covers a smaller range than does the mixer when both ranges are expressed as a percentage of frequency. At frequencies above 7000 kc. and with ordinary intermediate frequencies, the difference in percentage between the two tuning ranges is so small that it may be disregarded in receivers designed to cover only a small range, such as an amateur band.

A mixer and oscillator tuning arrangement in which a series tracking condenser is provided is shown in Figure 16. The value of the tracking condenser varies considerably with different intermediate frequencies and tuning ranges, capacities as low as .0001 μfd. being used at the lower tuning-range frequencies, and values up to .01 μfd. being used at the higher frequencies.

Bandspread Tuning. The frequency to which a receiver responds may be varied by changing the size of either the coils or the condensers in the tuning circuits, or both. In short-wave receivers a combination of both methods is usually employed, the coils being changed from one band to another, and variable condensers being used to tune the receiver across each band. In practical receivers, coils may be changed by one of two methods: a switch, controllable from the panel, may be used to switch coils of different sizes into the tuning circuits or, alternatively, coils of different sizes may be plugged manually into the receiver, the connection into the tuning circuits being made by suitable plugs on the coils. Where there are several "plug-in" coils for each band, they are sometimes arranged on a single mounting strip, allowing them all to be plugged in simultaneously.

In receivers using large tuning condensers to cover the short-wave spectrum with a minimum of coils, tuning is likely to be quite difficult, owing to the large frequency range covered by a small rotation of the variable condensers. To alleviate this condition, some method of slowing down the tuning rate, or *bandspreading*, must be used.

Quantitatively, bandspread is usually designated as being inversely proportional to the range covered. Thus, a *large* amount of bandspread indicates that a *small* frequency range is covered by the bandspread control. Conversely, a *small* amount of bandspread is taken to mean that a *large* frequency range is covered by the bandspread dial.

Types of Bandspread. Bandspreading systems are of two general types: electrical and mechanical. Mechanical systems are exemplified by high-ratio dials in which the tuning condensers rotate much more slowly than the dial knob. In this system there is often a separate scale or pointer either connected or geared to the dial knob to facilitate accurate dial readings. However, there is a limit to the amount of mechanical bandspread which can be obtained in an inexpensive dial before the speed-reduction unit develops backlash, which makes tuning difficult. To overcome this problem, most receivers employ a combination of both electrical and mechanical bandspread. In this system, a moderate reduction in the tuning rate is obtained in the dial, and the rest of the reduction obtained by *electrical bandspreading*.

Parallel Bandspread. In one form of electrical bandspread, two tuning condensers are used in parallel across each coil, one of rather high capacity to cover a large tuning range, and another of small capacity to cover a small range around the frequency to which the large condenser is set. These condensers are usually controlled by separate dials or knobs, the large condenser being known as the *bandsetting* condenser, and the smaller one being the bandspread condenser. Where there is more than one tuned circuit in the receiver, a bandsetting and a bandspread condenser are used across *each* coil, and all the condensers

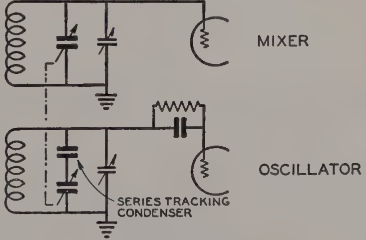


Figure 16.
OSCILLATOR SERIES TRACKING CONDENSER ARRANGEMENT.
The series condenser allows the oscillator to have a slower rate of capacity change than the mixer.

serving in each capacity are mechanically connected together, or *ganged,* thus allowing a single dial to be used for each purpose, even though there may be several tuned circuits.

Since the tuning range of a tuned circuit is proportional to the ratio of minimum to maximum capacity across it, a wide variation in the amount of bandspreading is made possible by a proper choice of the two capacities. The greater the capacity of the bandsetting condenser in proportion to the bandspread condenser, the greater will be the bandspread.

The bandspreading method described above is usually known as the *parallel* system. This system, as applied to a single tuned circuit, is diagrammed in Figure 17A. The large tuning, or bandsetting, condenser, C_F, usually has a maximum capacity of from 100 to 370 $\mu\mu$fd. C_B, the bandspread condenser, usually has a value of from 10 to 50 $\mu\mu$fd., depending upon the design of the receiver.

Dual-Rotor Bandspread. A special form of the parallel bandspread method is used in some manufactured tuning assemblies. In this system, a single set of stationary plates (stator) in the tuning condenser is acted upon by two separate rotors, one of large capacity for bandsetting and the other of small capacity for bandspread. Each rotor is operated by a separate dial. This system allows the bandsetting and bandspread functions to be combined in a single tuning-condenser unit. A variation of this method is sometimes used, in which the same dial is used for both bandsetting and bandspreading purposes, the change from one function to the other being accomplished by a "gear-shifting" mechanism built into the dial. The schematic of this bandspread system is shown in Figure 17B.

The parallel system of bandspreading has one major disadvantage, especially for amateur-band usage. This disadvantage lies in the fact that if the bandspreading condenser is made large enough to cover the lower-frequency amateur bands with optimum capacity being used across the coil in the bandsetting condenser, an extremely large bandsetting condenser is needed to give an equal amount of bandspread on the high-frequency bands. The high capacity across the coils reduces the impedance of the tuned circuits on the high-frequency bands.

Parallel-Bandspread Calculations. The following formulas will be found useful in designing parallel-bandspread circuits:

$$C_F = \frac{C_B F_L^2}{F_H^2 - F_L^2}, \text{ where}$$

C_F=Capacity of "bandsetting" condenser (μfd. or $\mu\mu$fd.)

C_B=Capacity range of bandspread condenser (same units as C_F)

F_L=Low-frequency end of tuning range (kc. or Mc.)

F_H=High-frequency end of tuning range (same units as F_L)

Where it is desired to know the number of turns to wind on a coil:

$$N = \sqrt{\frac{380{,}000\,(D + 3L)\,(F_H^2 - F_L^2)}{D^2\, C_B\, F_H^2\, F_L^2}}, \text{ where}$$

N=Number of turns

D=Diameter of coil, in inches

L=Length of coil, in inches

F_H=High-frequency end of tuning range, in megacycles

F_L=Low-frequency end of tuning range, in megacycles

C_B=Capacity range of bandspread condenser, in $\mu\mu$fd.

In both the above formulas C_B represents the amount of capacity variation supplied by the bandspread condenser. In well-designed midget condensers, the variation will approach the rated maximum capacity, and the maximum capacity may be used for C_B without serious error. In the first formula, the result C_F, will include all fixed capacities across the circuit, including the input capacity of the tube, stray capacity to ground, and the minimum capacity of the bandspread condenser.

For the special case where the bandspread circuit is to cover an amateur band, the following formulas apply:

$$C_F = \frac{C_B}{K}, \text{ and}$$

$$N = \sqrt{\frac{380{,}000\,K\,(D + 3L)}{D^2\, C_B\, F_H^2}}, \text{ where}$$

C_F, C_B, N, D, F_H, and L have the same significance as in the preceding two corresponding formulas and K has the following values:

160-meter band (1700–2100 kc.), K=.526
80-meter band (3450–4050 kc.), K=.378
40-meter band (6950–7350 kc.), K=.1184
20-meter band (13,950–14,450 kc.), K=.0729
10-meter band (27,950–30,050 kc.), K=.1559

Note that the above values of K are for a range 50-kc. wider than each amateur band.

Series Bandspread. The circuit shown in Figure 17C is known as *series bandspread.* This circuit is really nothing more than the parallel bandspread circuit just described, but with the single bandspread tuning condenser replaced by two condensers, C_S and C_{BS}, in series. The object of the series connection is to allow a single bandspread condenser to be used for all tuning ranges, yet to make it possible to change the effective capacity of the bandspread condenser when bands are changed. Changing the effective value of the bandspread condenser from band to band eliminates the disadvantage of the parallel bandspread system in regard to securing an advantageous L/C ratio on all bands, at the same time having a full-dial bandspread. In this system, the bandspread condenser, C_{BS}, usually has a capacity of 100 to 150 $\mu\mu$fd., while the "series" bandsetting condenser, C_S, may have a capacity of 25 to 50 $\mu\mu$fd. C_F is the usual "trimmer" condenser, and serves to set the minimum capacity across the whole circuit. The principle upon which the circuit operates is that while the *minimum* capacity across the coil varies but little for any setting of C_S, the *maximum* capacity available may be varied considerably by varying C_S.

The following formulas apply to the series bandspread arrangement:

General—

$$C_S = \frac{C_{BS}\, C_F\, (F_H^2 - F_L^2)}{C_{BS}\, F_L^2 - C_F\, (F_H^2 - F_L^2)},$$

$$C_F = \frac{C_{BS}\, C_S\, F_L^2}{(C_{BS} + C_S)\, (F_H^2 - F_L^2)},$$

$$N = \sqrt{\frac{380{,}000 (D+3L) (F_H^2 - F_L^2) (C_{BS} + C_S)}{D^2\, C_{BS}\, C_S\, F_H^2\, F_L^2}},$$

where C_F, F_H, F_L, N, D, and L have the same significance as in the formulas for parallel bandspread. C_{BS} and C_S are the bandspread, or "tuning" series condenser and the bandsetting or "fixed" series condenser, respectively. In the formulas for C_S and C_F, the values of C_{BS}, C_S, and C_F must all be in the same units, as must F_H and F_L. In the formula for N, C_F, C_S and C_{BS} must be in $\mu\mu$fd.; F_H and F_L must be in megacycles; and D and L must be in inches.

Amateur Bands—

$$C_S = \frac{KC_{BS}\, C_F}{C_{BS} - KC_F},$$

$$C_F = \frac{C_{BS}\, C_S}{K\, (C_{BS} + C_S)},$$

$$N = \sqrt{\frac{380{,}000\, K\, (D+3L) (C_{BS} + C_S)}{D^2\, C_{BS}\, C_S\, F_H^2}}, \text{ where}$$

K has the values given with the formulas for parallel bandspread. Again, in the formulas for C_S and C_F, the condenser capacities must be in the same units, and in the formula for N, the capacities must be in $\mu\mu$fd., the coil dimensions in inches, and F_H in megacycles.

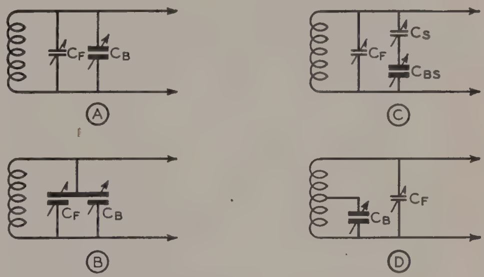


Figure 17.
BANDSPREAD CIRCUITS.
Parallel bandspread is shown at A and B, series bandspread at C, and tapped-coil bandspread at D. The operation of these circuits is discussed in the text.

One great disadvantage of series bandspread is the extreme non-linearity of tuning when the bandspread and "fixed" series condensers have widely different capacities. The net capacity across the coil changes rapidly with the change in bandspread capacity, when the bandspread condenser is near minimum capacity. Thus, as the bandspread condenser is rotated toward minimum capacity, the change in net capacity becomes more and more for each degree of angular rotation. Series bandspread thus causes the high frequency end of the band to be squeezed into a few dial degrees and the rest of the band to spread out across the rest of the dial range. Some of this difficulty can be eliminated by placing a fixed minimum capacity across the bandspread condenser to keep the minimum capacity substantially greater than the capacity of the series condenser, or by using a bandspread condenser with specially shaped plates. The first method has the disadvantage of requiring still another condenser in the tuned-circuit, while the second requires a tuning condenser ordinarily not available to the home constructor.

Tapped-Coil System. To allow equal bandspread on the amateur bands and still not use extremely high bandsetting capacities on the higher frequencies, the variation of the parallel system shown in Figure 17D is often employed. As the bandspread condenser, C_B, is connected across part of the coil, this method is usually known as the *tapped coil* system.

The effectiveness of the bandspread condenser in tuning the coil depends upon the amount of the coil included across the bandspread condenser terminals. As the number of turns between the bandspread condenser terminals is decreased, the amount of bandspread increases.

In most amateur-band receivers employing the tapped-coil system of bandspreading, a separate bandsetting condenser is permanently connected across each coil. These condensers are either mounted within the coils, in the plug-in-coil system, or alongside the coils in the bandswitching system.

The tapped-coil bandspread method is quite widely used in modern amateur-band receivers, especially in home constructed sets. Its principal advantage is that it allows equal bandspread, to any degree desired, over several amateur bands. Another advantage is that it facilitates accurate tracking in ganged tuning circuits; the coil taps are adjusted until the circuits track identically.

Best results with the tapped-coil system will be obtained when C_B is made just large enough to tune the widest band when connected completely across a suitable coil, and then tapping C_B down the required amount on the narrower bands. (By "widest band" is meant the widest in terms of percentage, not kilocycles. The widest amateur band, thus, is the 160-meter band, even though it is next to the narrowest in terms of actual kilocycles.)

Calculating the correct point for the location of the tap in the tapped-coil system is rather complicated, and for this reason, tapped-coil data is given in the form of design charts for each amateur band in Figure 18. When either the inductance of the coil or the capacity required for resonance at the high-frequency end of the band is known, the charts will give the proper location of the bandspread tap with good accuracy.

As with the preceding formulas for series and parallel bandspread, the tapped-coil charts apply only to circuits tuned to the signal frequency; i.e., they are not accurately applicable to superheterodyne oscillator coils. The lower the i.f. channel frequency and the higher the signal frequency the greater will be their accuracy for oscillator coil purposes. When the intermediate frequency is 465 kc. or lower, the curves for the three higher frequency bands will serve with acceptable accuracy for locating the oscillator coil bandspread tap.

Circuit Capacity. In this book and in other radio literature, mention is sometimes made of "stray" or *circuit capacity.* This capacity is in the usual sense defined as the capacity remaining across a coil when all the tuning, bandspread, and padding condensers across the circuit are at their minimum capacity setting. Circuit capacity can be attributed to two general sources. One source, which is fixed for any particular type of tube, is that due to the "cold" input capacitance of the tube when its cathode is not heated. The input capacitance varies somewhat from the fixed value when the tube is in actual operation. Such factors as plate load impedance, grid bias, and frequency will cause a change in input capacitance. However, in all except the extremely high-transconductance tubes, the published measured input capacitance is quite close to the effective value. In the high-transconductance types, however, the effective capacitance does vary considerably from the published figures, under different operating conditions.

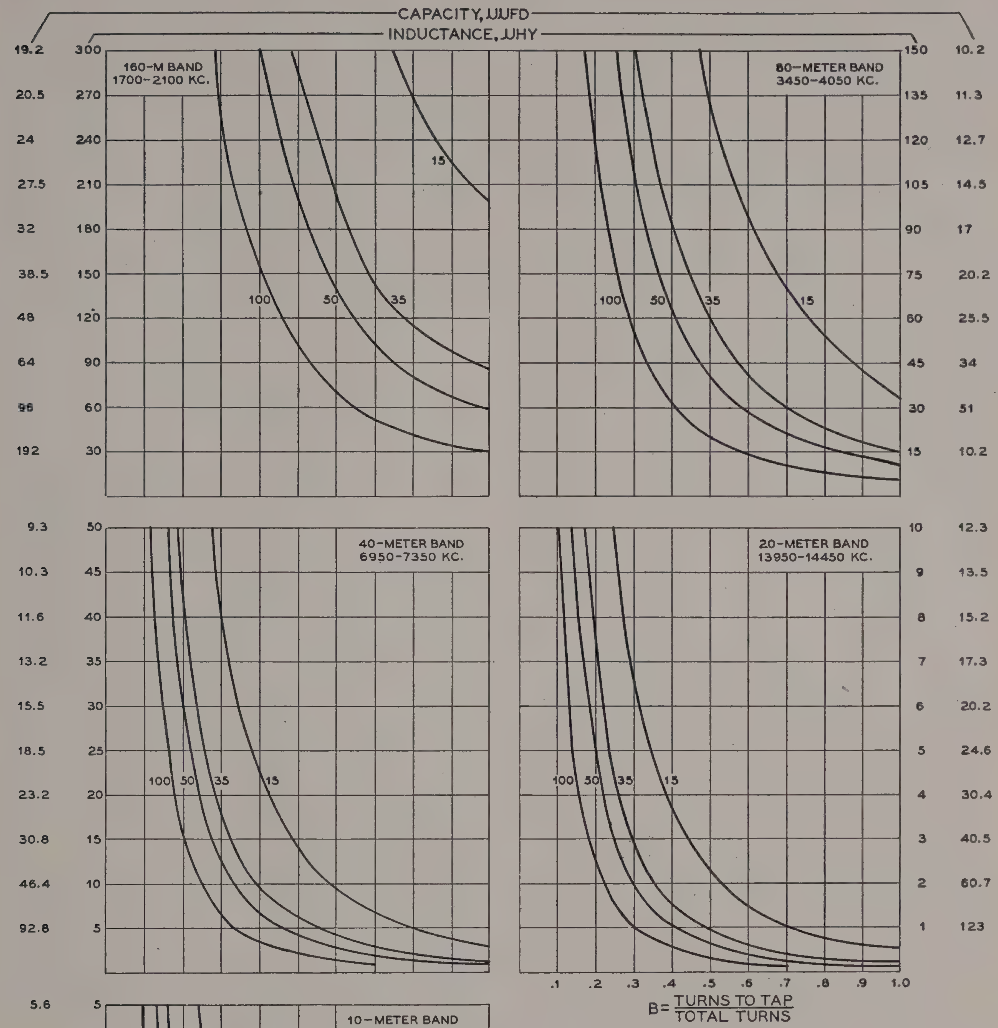
The second source of circuit capacity, and that which is more easily controllable, is that contributed by the minimum capacity of the variable condensers across the circuit and that due to capacity between the wiring and ground. In well-designed high-frequency receivers, every effort is made to keep this portion of the circuit capacity at a minimum, since a large capacity reduces the tuning range available with a given coil and prevents a good L/C ratio, and consequently a high-impedance tuned circuit, from being obtained.

Typical values of circuit capacity may run from 10 to 75 $\mu\mu$fd. in high-frequency receivers, the first figure representing concentric-line receivers with acorn or miniature tubes and extremely small tuning condensers, and the latter representing all-wave sets with bandswitching, large tuning condensers, and conventional tubes.

I.F. Tuned Circuits

All i.f. amplifiers employ bandpass circuits of some sort. A bandpass circuit is exactly what the name implies—a circuit for passing a band of frequencies. Bandpass arrangements can be designed for almost any degree of selectivity, the type used in any particular application depending upon the use to which the i.f. amplifier is to be put.

Bandpass Circuits. Bandpass circuits consist essentially of two or more tuned circuits and some method of coupling the tuned circuits together. Some representative arrange-



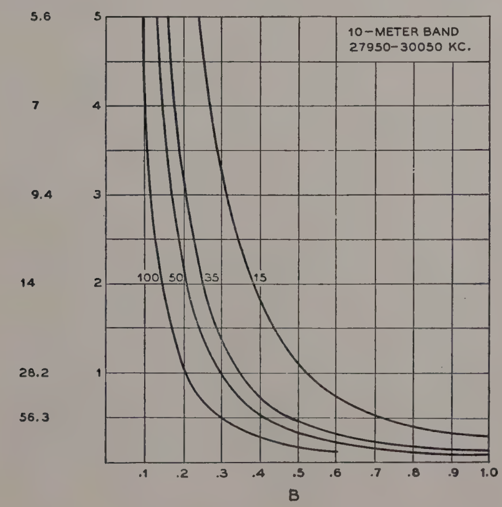


Figure 18.

TAPPED-COIL BANDSPREAD CHARTS.

To use the charts, locate the coil inductance in the columns at the right or left opposite the band in question. Reading to the right or left opposite the coil inductance, the proper location of the bandspread tap "B" will be found below the intersection of the inductance and the curve marked with the bandspread condenser capacity. The capacities given in the vertical columns are those required to resonate a coil of the indicated inductance at the high-frequency end of the tuning range with the bandspread condenser disconnected from the coil. This capacity will include the parallel trimmer condenser and all stray capacity across the coil. Greatest accuracy in using the charts will probably be obtained by calculating the coil inductance directly from the formula given on page 35. The charts assume that the coil length and diameter are equal, but will be acceptably accurate for moderate deviations from this d/l ratio.

ments are shown in Figure 19. The circuit shown at A is the conventional i.f. transformer, with the coupling, M, between the tuned circuits being provided by inductive coupling from one coil to the other. As the coupling is increased, the selectivity curve becomes less peaked, and when a condition known as "critical coupling" is reached, the top of the curve begins to flatten out. When the coupling is increased still more, a dip occurs in the top of the curve. The windings for this type of i.f. transformer, as well as most others, nearly always consist of small, flat universal-wound pies mounted either on a piece of dowel to provide an air core or on powdered-iron impregnated bakelite for "iron core" i.f. transformers. The iron-core transformers generally have somewhat more gain and better selectivity than equivalent air-core units between 175 and 2000 kc.

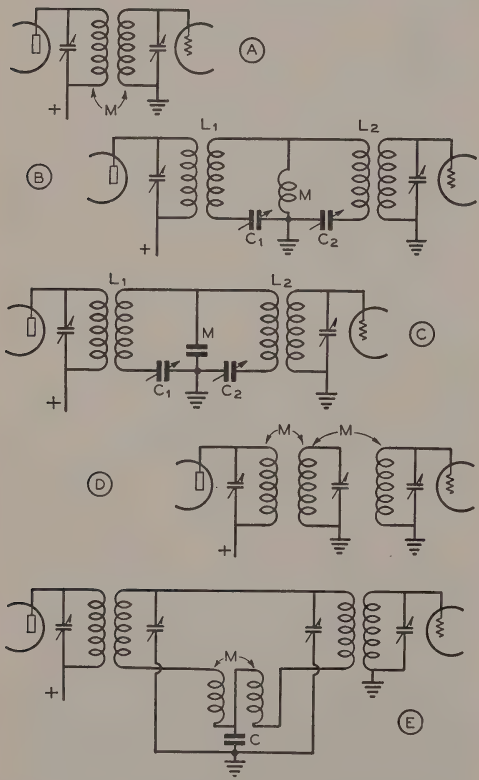


Figure 19.
I.F. AMPLIFIER BAND-PASS CIRCUITS.
The ordinary i.f. transformer circuit is shown at A. Other circuits are intended to give a straight-sided, flat-topped selectivity characteristic to the i.f. amplifier.

The circuits shown at B and C are quite similar. Their only difference is the type of mutual coupling used, an inductance being used at B and a capacitance at C. The operation of both circuits is similar. Three resonant circuits are formed by the components. In B, for example, one resonant circuit is formed by L_1, C_1, C_2, and L_2 all in series. The frequency of this resonant circuit is just the same as that of a single one of the coils and condensers, since the coils and condensers are similar in both sides of the circuit, and the resonant frequency of the two condensers and the two coils all in series is the same as that of a single coil and condenser. The second resonant frequency of the complete circuit is determined by the characteristics of each half of the circuit containing the mutual coupling device. In B, this second frequency will be lower than the first, since the resonant frequency of L_1, C_1 and the inductance, M, or L_2, C_2 and M is lower than that of a single coil and condenser, due to the inductance of M being added to the circuit. The opposite effect takes place at C, where the common coupling impedance is a condenser. Thus, at C the second resonant frequency is higher than the first. In either case, however, the circuit has two resonant frequencies, resulting in a flat-topped selectivity curve. The width of the top of the curve is controlled by the reactance of the mutual coupling component. As this reactance is increased (inductance made greater, capacity made smaller), the two resonant frequencies become farther apart and the curve is broadened.

The circuit of Figure 19D is often used where a fairly high degree of bandpass action is required and the number of i.f. transformers used must be kept at a minimum. In this circuit there is inductive coupling between the center coil and each of the outer coils. The result of this arrangement is that the center coil acts as a sharply tuned coupler between the other two. A signal somewhat off the resonant frequency of the transformer will not induce as much voltage in the center coil as will a signal of the correct frequency. When a smaller voltage is induced in the center coil, it in turn transfers a still smaller voltage to the output coil. In other words, the effective coupling between the outer coils increases as the resonant frequency is approached, and remains nearly constant over a small range and then decreases again as the resonant band is passed.

Another very satisfactory bandpass arrangement, which gives a very straight-sided, flat-topped curve, is the negative-mutual arrangement shown at E. Energy is transferred between the input and output circuits in this arrangement by both the negative-mutual coils,

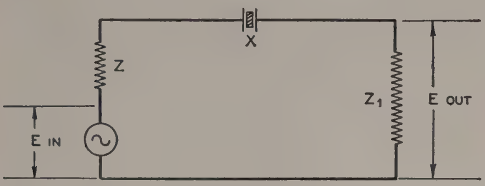


Figure 20.
CRYSTAL FILTER EQUIVALENT CIRCUIT.
With a constant input voltage, the r.f. voltage developed across Z_1 depends upon the impedances of Z, X, and Z_1.

M, and the common capacitive reactance, C. The negative-mutual coils are interwound on the same form, and connected "backward," as shown.

Crystal Filters. The selectivity of the intermediate-frequency amplifier may be increased greatly through the use of an extremely high Q piezo-electric series resonant circuit. The piezo-electric quartz crystal, together with its coupling arrangement, is generally known as a *crystal filter.* The electrical equivalent of the basic crystal filter circuit is shown in Figure 20, while the electrical equivalent of the crystal itself is shown in Figure 21.

At its resonant frequency, the crystal, X, is equivalent to a small resistance, and thus at this frequency the current flowing through the circuit, Z, X, Z_1 reaches a maximum, and the output voltage E_{out} is also at its maximum value. At frequencies slightly off resonance, the crystal impedance becomes quite high and the current flowing through the circuit, and consequently the voltage E_{out} developed across Z_1, drops to a low value. It is the ratio of E_{out} at resonance to this voltage, at frequencies away from resonance, that determines the selectivity characteristic of the crystal filter. This ratio may be shown to depend upon the values of the impedances Z and Z_1. These impedances remain nearly constant for frequencies near crystal resonance, and the selectivity of the filter circuit as a whole may be altered by changing their resonant frequency values. The variable selectivity crystal filter circuits quite often used in communications superheterodynes operate on this principle.

Practical Filters. In practical crystal filters it is necessary to balance out the capacity across the crystal holder (C_1 in Figure 21) to prevent by-passing around the crystal of undesired signals off the crystal resonant frequency. The balancing is done by a *phasing* circuit which takes out-of-phase voltage from a balanced input circuit and passes it to the output side of the crystal in proper phase to neutralize that passed through the holder capacity. A representative practical filter arrangement is shown in Figure 22. The phasing condenser is indicated in the diagram by PC. The balanced input circuit may be obtained either through the use of a split-stator condenser as shown, or by the use of a center-tapped input coil.

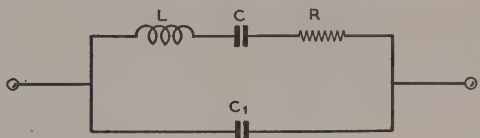


Figure 21.
CRYSTAL EQUIVALENT.
The crystal is equivalent to a very large inductance in series with a very small resistor and condenser.

Variable-Selectivity Filters. In the circuit of Figure 22, the selectivity is *minimum* with the crystal input circuit tuned to resonance, since at resonance the input circuit is a pure resistance effectively in series with the voltage applied to the crystal. As the input circuit is detuned from resonance, however, the resistive component of the input impedance decreases, and the selectivity becomes greater. In this circuit, the output from the crystal filter is tapped down on the i.f. stage grid winding to provide a better match and lower the impedance in series with the crystal.

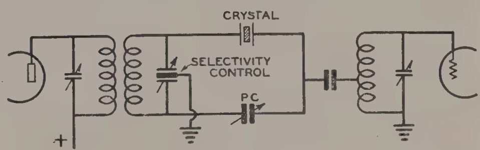


Figure 22.
VARIABLE-SELECTIVITY CRYSTAL CIRCUIT.
In this circuit the selectivity is at a minimum when the input circuit is tuned to resonance.

The circuit shown in Figure 23 also achieves variable selectivity by adding an impedance in series with the crystal circuit. In this case, the variable impedance is in series with the crystal output circuit. The impedance of the output tuned circuit is varied by varying the Q. As the Q is reduced (by adding resistance in series with the coil), the impedance decreases and the selectivity becomes greater. The input circuit impedance is made low by using a non-resonant secondary on the input transformer.

A variation of the circuit shown at Figure 23

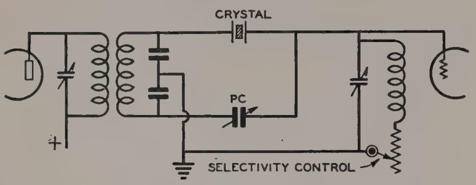


Figure 23.
WIDE-RANGE VARIABLE-SELECTIVITY CRYSTAL FILTER.
As the impedance of the crystal output circuit is lowered, by inserting more resistance in series with the coil, the selectivity of the filter circuit is improved.

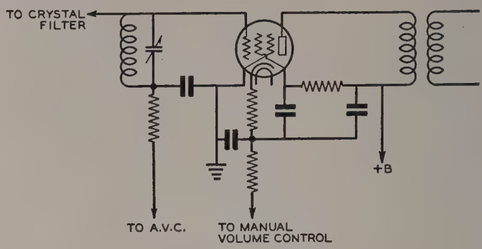


Figure 24.
DEGENERATIVE I.F. STAGE.
Degeneration in the i.f. stage following the crystal filter is desirable to avoid input capacity changes when the gain is varied.

consists of placing the variable resistance across the coil and condenser, rather than in series with them. The result of adding the resistor is a reduction of the output impedance, and an increase in selectivity. The circuit behaves oppositely to that of Figure 23, however; as the resistance is lowered the selectivity becomes greater. Still another variation of Figure 23 is to use the tuning condenser across the output coil to vary the output impedance. As the output circuit is detuned from resonance, its impedance is lowered, and the selectivity increases. Sometimes a set of fixed condensers and a multipoint switch are used to give step-by-step variation of the output circuit tuning, and thus of the crystal filter selectivity.

Interference Rejection. The crystal filter phasing condenser can be adjusted so that parallel resonance between it and the crystal causes a sharp dip in the response curve at some desired point, such as 2 kc. from the desired signal peak. This effect can be utilized to eliminate completely the unwanted sideband 1 kc. away from zero beat for c.w. reception. The b.f.o. then provides a true single signal effect, that is, a single beat frequency note. This effectively increases the number of c.w. channels that can be used in any shortwave band.

Reducing Input Capacity Variations. As the previous discussion on crystal filters has indicated, the selectivity of the crystal filter can be altered by changing the impedance of the crystal output circuit. Since the impedance at crystal frequency of the output circuit can be varied by detuning it as well as by varying its Q, it is important that the input capacity of the tube following the filter remain constant when the gain of this stage is varied. The input capacity may be stabilized with respect to changes in the tube's amplification by employing a small amount of degeneration, as illustrated in Figure 24. The amount of degeneration which can be used will depend upon the amount of gain which can be sacrificed in the i.f. stage following the crystal filter. Values for R will ordinarily fall between one-third and two-thirds of the total resistance in the cathode circuit, exclusive of the manual gain control.

Detector, Audio, and Control Circuits

Detectors. Second detectors for use in superheterodynes are usually of the diode, plate, or infinite impedance types, which were described in detail in Chapter 3. Occasionally, grid-leak detectors are used in receivers using one i.f. stage or none at all, when the second detector is regenerative.

Diodes are the most popular second detectors because they allow a simple method of obtaining automatic volume control to be used. Diodes load the tuned circuit to which they are connected, however, and thus reduce the selectivity slightly. Special i.f. transformers are used for the purpose of providing a low-impedance input circuit to the diode detector.

Automatic Volume Control. The elements of an automatic volume control (a.v.c.) system are shown in Figure 25. A dual-diode tube is used as a combination diode detector and a.v.c. rectifier. The left-hand diode operates as a simple rectifier in the manner described earlier in this chapter. Audio voltage, superimposed on a d.c. voltage, appears across the 500,000-ohm potentiometer (the volume control) and the .0001-μfd. condenser, and is passed on to the audio amplifier. The right-hand diode receives signal voltage directly from the primary of the last i.f. amplifier, and it acts as the a.v.c rectifier. The pulsating d.c. voltage across the 1-megohm a.v.c.-diode load resistor is filtered by a 500,000-ohm resistor and a .05-μfd. condenser, and applied as bias to the grids of the r.f. and i.f. amplifier tubes; an increase or decrease in signal

strength will cause a corresponding increase or decrease in a.v.c. bias voltage, and thus the gain of the receiver is automatically adjusted to compensate for changes in signal strength.

By disassociating the a.v.c. and detecting functions through using separate diodes, as shown, most of the ill effects of a.c. shunt loading on the detector diode are avoided. This type of loading causes serious distortion, and the additional components required to eliminate it are well worth their cost. Even with the circuit shown, a.c. loading can occur unless a very high (5 megohms, or more) value of grid resistor is used in the following audio amplifier stage. An alternative to using a high value of grid resistor is to connect the potentiometer arm directly to the audio stage grid, without an intervening coupling condenser. This method allows the d.c. voltage across the detector diode load resistor to be used as grid bias for the audio stage. Since the amount of bias will vary with changes in signal strength and changes in the setting of the volume control, it is essential that the audio stage be resistance coupled and have a high value of plate resistance when this method is used.

An a.v.c. circuit which may be added to a receiver not so equipped is shown in Figure 26. In this circuit, the pentode section of a duplex-diode-pentode is used as a resistance coupled i.f. amplifier which receives its excitation from the detector grid circuit. The output from the pentode is applied to the two diodes in parallel, through a coupling condenser, and the rectified voltage across the diode load resistor is used as a.v.c. bias.

A.V.C. in B.F.O.-Equipped Receivers. In receivers having a beat-frequency oscillator for the reception of radiotelegraph signals, the use of a.v.c. can result in a great loss in sensitivity when the b.f.o. is switched on. This is because the beat oscillator output acts exactly like a strong received signal, and causes the a.v.c. circuit to put high bias on the r.f. and i.f. stages, thus greatly reducing the receiver's sensitivity. Due to the above effect, it is necessary to provide a method of making the a.v.c. circuit inoperative when the b.f.o. is being used. The simplest method of eliminating the a.v.c. action is to short the a.v.c. line to ground when the b.f.o. is turned on. A two-circuit switch may be used for the dual purpose of turning on the beat oscillator and shorting out the a.v.c., if desired.

Signal Strength Indicators. Visual means for determining whether or not the receiver is

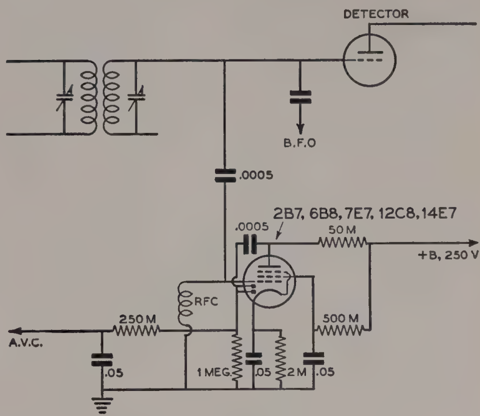


Figure 26.
A.V.C. CIRCUIT FOR ANY SUPERHETERODYNE.
This circuit may be added to a receiver not equipped with a.v.c. The duo-diode-pentode acts as an a.v.c. amplifier and diode rectifier.

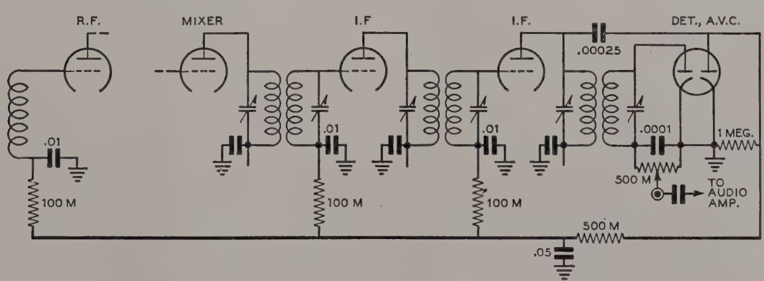


Figure 25.
DOUBLE-DIODE DETECTOR-A. V. C. CIRCUIT.
Any of the ordinary small dual-diode tubes may be used in this circuit. The left-hand diode serves as the detector, while the right-hand section operates as an a.v.c. rectifier. Using separate diodes for the detector and a.v.c. functions helps to improve the audio fidelity.

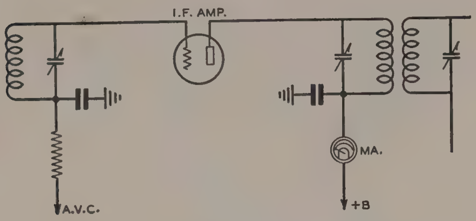


Figure 27.
USING A LOW-RANGE MILLIAMMETER AS A TUNING OR SIGNAL STRENGTH INDICATOR.
The plate current to an i.f. stage varies as the a.v.c. bias changes. A 0-10 d.c. milliammeter will serve in most cases. The meter reads "backwards" in this circuit, strong signals causing the current to decrease more than weak ones.

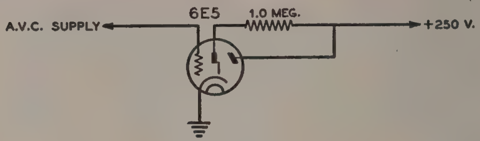


Figure 28.
ELECTRON-RAY TUNING INDICATOR.
Other "eye" tubes such as the 6U5 and 6AB5 may also be used in this circuit.

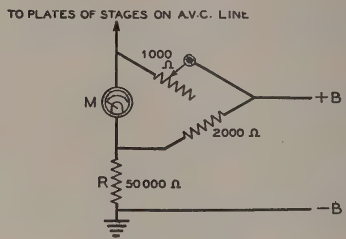


Figure 29.
FORWARD-READING SIGNAL STRENGTH METER CIRCUIT.
Placing the meter in a bridge circuit allows it to read in a "forward" direction in respect to signal strength. The meter is usually a 0-1 milliammeter.

properly tuned, as well as an indication of the relative signal strength, are both provided by means of *tuning indicators* of the meter or vacuum-tube types. Direct current milliammeters can be connected in the plate-supply circuit of an r.f. amplifier, as shown in Figure 27, so that the change in plate current, due to the a.v.c. voltage which is supplied to that tube, will indicate proper tuning. Sometimes these d.c. meters are built in such a manner as to produce a shadow of varying width. Vacuum-tube tuning indicators are designed so that an electron-ray "eye" pattern changes its size when the input circuit of the tube is connected across all or part of the a.v.c. voltage. The basic circuit for this type of indicator is illustrated in Figure 28.

Unfortunately, when an ordinary meter is used in the plate circuit of a stage, for the purpose of indicating signal strength, the meter reads backward with respect to strength. This is because increased a.v.c. bias on stronger signals causes lower plate current through the meter. For this reason, special meters which indicate zero at the right-hand end of the scale are often used for signal strength indicators in this type of circuit. Alternatively, the meter may be mounted upside down, so that the needle moves toward the right with increased signal strength.

A circuit which allows an ordinary meter to be used, and which gives conventional right-hand movement of the needle with increased signal strength, is shown in Figure 29. The plate (or plate and screen) current to the stages receiving a.v.c. bias is fed through one-half of a bridge network. The meter, M, is usually a 0–1 milliammeter. The resistor values shown are average ones; it may be necessary to change them slightly, depending upon the number of stages drawing current through the network. Using a lower value at R will give greater "swing" for a given signal strength, while larger values will reduce the swing. The variable 1000-ohm resistor is used to set the meter for minimum indication when no signal is being received.

Beat-Frequency Oscillators. The beat-frequency oscillator, usually called the *b.f.o.*, is a necessary adjunct for reception of c.w. telegraph signals on superheterodynes which do not use regenerative detectors. The oscillator is coupled into the second detector circuit and supplies a signal of nearly the same frequency as that of the desired signal from the i.f. amplifier. If the i.f. amplifier is tuned to 465 kc., for example, the b.f.o. is tuned to approximately 464 or 466 kc. in order to produce a 1000-cycle beat note in the output of the second detector of the receiver. The carrier signal would otherwise be inaudible. The b.f.o. is not used for voice reception, except as an aid in searching for weak stations.

The b.f.o. input to the second detector need only be sufficient to give a good beat note on an average signal. Too much coupling into the second detector will give an excessively high hiss level, masking weak signals by the high noise background.

A method of manually adjusting the b.f.o. output to correspond with the strength of received signals is shown in Figure 30. A variable b.f.o. output control of this sort is a useful adjunct to any superheterodyne, since it allows sufficient b.f.o. output to be obtained to give a "beat" with strong signals and at the same time permits the b.f.o. output, and consequently the hiss, to be reduced when attempting to receive weak signals. The circuit shown is somewhat better than those in which one of the electrode voltages on the b.f.o. tube is changed, as the latter usually change the frequency of the b.f.o. at the same time they change the strength, making it necessary to reset the trimmer each time the output is adjusted.

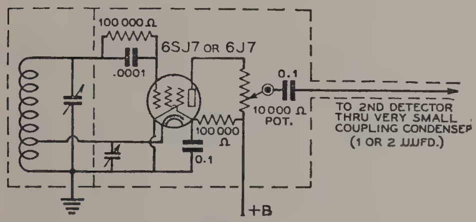


Figure 30.
VARIABLE-OUTPUT B.F.O. CIRCUIT.
Being able to vary the output of the b.f.o. is sometimes helpful when receiving weak signals.

Noise Suppression

The problem of noise suppression confronts the listener who is located in such places where interference from power lines, electrical appliances, and automobile ignition systems is troublesome. This noise is often of such intensity as to swamp out signals from desired stations.

There are three principal methods for reducing this noise:

(1) A.c. line filters at the source of interference, if the noise is created by an electrical appliance.
(2) Noise-balancing circuits for the reduction of power-leak interference.
(3) Noise-limiting circuits for the reduction, in the receiver itself, of interference of the type caused by automobile ignition systems.

Power Line Filters. Numerous household appliances, such as electric mixers, heating pads, vacuum sweepers, refrigerators, oil burners, sewing machines, doorbells, etc., create an interference of an intermittent nature. The

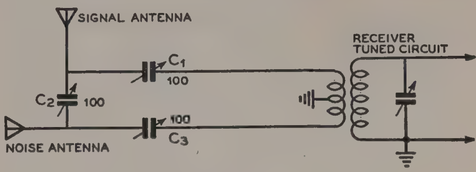


Figure 31.
JONES NOISE-BALANCING CIRCUIT.
This circuit, when properly adjusted, reduces the intensity of power-leak and similar interference.

insertion of a line filter near the source of interference often will effect a complete cure. Filters for small appliances can consist of a 0.1-μfd. condenser connected across the 110-volt a.c. line. Two condensers in series across the line, with the midpoint connected to ground, can be used in conjunction with ultra-violet ray machines, refrigerators, oil burner furnaces, and other more stubborn offenders. In severe cases of interference, additional filters in the form of heavy-duty r.f. choke coils must be connected in series with the 110-volt a.c. line on both sides of the line.

Noise Balancing. Power line noise interference can be greatly reduced by the installation of a *noise-balancing* circuit ahead of the receiver, as shown in Figure 31. The noise-balancing circuit adds the noise components from a separate noise antenna in such a manner that this noise antenna will buck the noise picked up by the regular receiving antenna. The noise antenna can consist of a connection to one side of the a.c. line, in some cases, while at other times an additional wire, 20 to 50 feet in length, can be run parallel to the a.c. house supply line. The noise antenna should pick up as much noise as possible in comparison with the amount of signal pickup. The regular receiving antenna should be a good-sized outdoor antenna, so that the signal-to-noise ratio will be as high as possible. When the noise components are balanced out in the circuit ahead of the receiver, the signals will not be appreciably attenuated.

This type of noise balancing is not a simple process; it requires a bit of experimentation in order to obtain good results. However, when proper adjustments have been made, it is possible to reduce the power leak noise from 3 to 5 "S" points without reducing the signal strength more than one S point, and in some cases there will be no reduction in signal strength whatsoever. This means that fairly weak signals can be received through terrific power leak interference. Hash type interference from electrical appliances can be re-

duced to a very low value by means of the same circuits.

The coil should be center-tapped and connected to the receiver ground connection in most cases. The pickup coil consists of 4 turns of hookup wire 2 inches in diameter, which can be slipped over the first r.f. tuned coil in most radio receivers. A 2-turn coil is more appropriate for 10- and 20-meter operation, though the 4-turn coil is suitable if care is taken in adjusting the condensers to avoid 10-meter resonance (unless very loose inductive coupling is used).

Adjustment of C_1 will generally allow a noise balance to be obtained when varying C_2 and C_3. One antenna, then the other, can be removed to check for noise in the receiver. When properly balanced, the usual power line buzz can be balanced down nearly to zero without attenuating the desired signal more than 50 per cent. This may result in the reception of an intelligible distant signal through extremely bad power line noise. Sometimes an incorrect adjustment will result in balancing out the signal as well as the noise. A good high antenna for signal reception will ordinarily overcome this effect.

With this circuit, some readjustment is necessary from band to band in the short-wave spectrum; noise-balancing systems require a good deal of patience and experimenting at each particular receiving location.

Noise-Limiting Circuits. Several different noise-limiting circuits have become popular. These circuits are beneficial in overcoming automobile ignition interference. They operate on the principle that each individual noise pulse is of very short duration, yet of extremely high amplitude. The popping or clicking type of noise from electrical ignition systems may produce a signal having a peak value ten to twenty times as great as the incoming radio signal.

As the duration of this type of noise peak is short, the receiver can be made inoperative during the noise peak without the human ear detecting the total loss of signal. Some noise limiters, or eliminators, actually *punch a hole* in the signal, while others merely *limit* the maximum peak signal which reaches the headphones or loudspeaker.

The noise peak is of such short duration that it would not be objectionable except for the fact that it produces an overloading effect on the receiver, which increases its time constant. A sharp voltage peak will give a kick to the diaphragm of the headphones or speaker, and the momentum or inertia keeps the diaphragm in motion until the dampening of the diaphragm stops it. This movement produces a popping sound which may completely obliterate the desired signal. If the noise peak can be limited to an amplitude equal to that of the desired signal, the resulting interference is practically negligible.

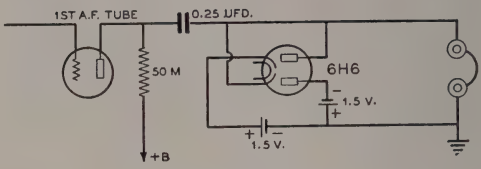


Figure 32.
A.F. NOISE LIMITER.
A limiter such as this is effective in reducing short-duration noise pulses, such as automobile ignition interference.

A.F. Peak Limiters. Remarkably good noise suppression can be obtained in the audio amplifier of a radio receiver by using a delayed push-pull diode suppressor.

The circuit in Figure 32 can be used to describe the operation of this general type of noise suppressor or limiter. Each diode works on opposite noise voltages; that is, both sides of the noise voltage (+ and − portions of the a.c. components), are applied to diodes which short-circuit the load whenever the applied voltage is greater than the delay voltage. The delay bias voltage prevents diode current from flowing for low-level audio voltages, and so the noise circuit has no effect on the desired signals except during the short interval of noise peaks. This interval is usually so short that the human ear will not notice a drop in signal during the small time that the load (headphones) is short-circuited by the diodes.

Delay bias voltage of 1½ volts from a small flashlight cell will allow any signal voltage which has a peak of less than about 1½ volts to operate the headphones. Noise peaks often have values of from 5 to 20 times as great as the desired signal; so these peaks operate the diodes, causing current to flow and a sudden drop in impedance across the headphones.

Diodes have nearly infinite impedance when no diode current is flowing; however, as soon as current starts, the impedance will drop to a very few hundred ohms, which tends to damp out or short circuit the audio output. The final result is that the noise level from automobile ignition is limited to values no greater than the desired signal. This is low enough to cause no trouble in understanding the voice or c.w. signals.

A push-pull diode circuit is necessary because the noise peaks are of an a.c. nature and are not symmetrical with respect to the zero

a.c. voltage reference level. The negative peaks may be greater than the positive peaks, depending on the bias and overload characteristics of the audio amplifier tube. If a single diode is used, only the positive (or negative) peaks could be suppressed. In Figure 29 the two bias dry-cells are arranged to place a negative bias on each diode plate of 1½ volts. A positive noise voltage peak at the plate of the audio amplifier tube will overcome this negative bias on the top diode plate and cause diode current to flow and lower the impedance. A negative noise voltage peak will overcome the positive bias on the other diode cathode and cause this diode to act as a noise suppressor. A positive bias on the cathode is the same as a negative bias on the diode plate. The 6H6 has two separate cathodes and plates, hence lends itself readily to the simple circuit illustrated in Figure 32.

Circuits of this type are very effective for short-pulse noise elimination because they tend to punch a hole in the signal for the duration of a strong noise voltage peak. A peak that will cause a loudspeaker or headphones to rattle with a loud pop will be reduced to a faint pop by the noise-limiting system. The delay bias prevents any attenuation of the desired signal as long as the signal voltage is less than the bias.

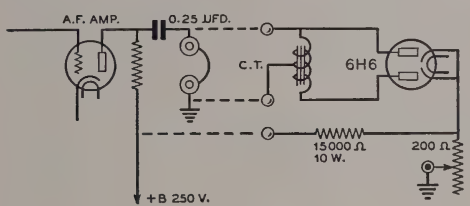


Figure 33.
ADJUSTABLE NOISE LIMITER.
With this circuit the bias on the limiter diodes is adjustable for different noise levels. The center-tapped choke may be the primary of a small pentode output transformer.

With this type of noise limiter it is possible to adjust the audio or sensitivity gain controls so that the auto ignition noise seems to drop out, leaving only the desired signal with a small amount of distortion. Lower gain settings will allow some noise to get through but will eliminate audio distortion on voice or music reception. At high levels, the speech or music peaks will be attenuated whenever they exceed the d.c. delay bias voltage. Faint ignition rattle will always be audible in the background with any noise-suppressor circuit, since

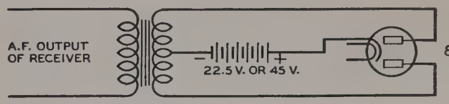


Figure 34.
NOISE LIMITER FOR USE WITH LOUDSPEAKER.
The high bias on this dual-diode noise limiter allows it to be used on high-level audio stages.

some noise peaks are too small to operate the systems, yet are still audible as a weak rattle or series of pops in the headphones.

Figures 33 and 34 show two noise-limiter circuits which can be used as separate units for connection to any receiver. The unit shown in Figure 33 can be connected across any headphone output as long as there is no direct current flowing through the phones. A blocking condenser can be connected in series with it if necessary, though better noise suppression results when the blocking condenser is in series with the plate lead to the headphones. Delay bias is obtained from the plus B supply through a 15,000-ohm 10-watt resistor and a 200-ohm wire-wound variable resistor. The cathode or cathodes are made a volt or two positive with respect to ground and minus B connection.

The diode plates are connected through a center-tapped low resistance choke to ground as far as bias voltage is concerned. Any push-pull to voice coil output transformer can be used for the center-tapped choke in Figure 33. The secondary can be left open. The delay bias is adjustable from 0 up to about 3 volts and once set for some noise level, can be left in that position.

The unit illustrated in Figure 34 can be connected across any audio amplifier stage, even the output stage which drives a loudspeaker. Any bias from 1½ to 90 volts or more can be connected in series with the center tap and the 6Z4 tube cathode. The higher values of delay bias would be needed for high output levels from the loudspeaker. Generally, 22½- to 45-volts bias will allow enough delay to allow moderate room volume reception of the desired voice signals without leveling off and distortion. The bias should be as low as possible, without distortion, in order to obtain effective noise suppression.

Second-Detector Noise Limiters. There are numerous arrangements for noise limiting in the second detector circuit. Tests conducted with a great many of these circuits have indicated that the ones shown in Figures 35, 36 and 37 are the most practical and desirable

for use in amateur communications receivers. The noise-silencing action of these limiters is obtained either by shorting the noise pulses to ground, or by opening an "electronic switch" in series with the audio current on each noise pulse. The circuit of Figure 35 is an example of the first method, while those of Figures 36 and 37 are of the latter type.

The *Dickert* noise limiter circuit shown in Figure 35 makes use of a diode detector and a small class B triode such as the 6A6, 6N7, or 79 as the noise limiter tube. The latter tubes are used because, at zero or negative grid voltage and a small amount of plate potential, they draw very little plate current.

Under normal operation, with a received carrier, the grid of the 6N7 is biased negatively by an amount slightly less than half the rectified carrier voltage. This means that for modulation percentages up to nearly 100 per cent, the resistance of the 6N7 will remain very high due to its grid always remaining negative with respect to the cathode. Note also that the grid is supplied with d.c. through a filter circuit with a comparatively high time constant, so that the actual grid potential varies but very slowly with changing external conditions.

With the reception of a noise pulse the cathode of the 6N7 is instantaneously driven highly negative, while the control grid maintains the moderate carrier-level bias due to the time constant of the filter feeding it. Another way of stating that the cathode goes negative with respect to the grid is, of course, to say that the control grid is driven positive.

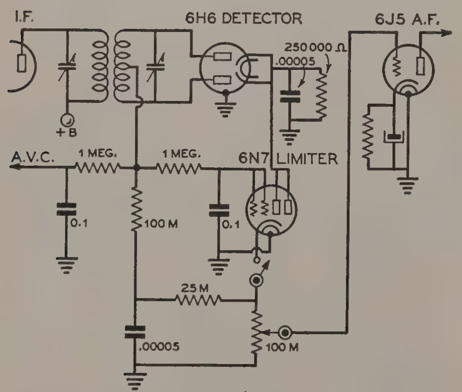


Figure 35.

DICKERT AUTOMATIC LIMITER.

This limiter will automatically adjust itself to various amounts of carrier strength. The recommended values of components are shown.

Also, at the same time that the control grid goes positive, the same noise pulse drives the plate of the 6N7 more positive due to the common resistance between it and the cathode of the detector, and ground. This of course means that the current due to the noise pulse flows almost entirely between the cathode and plate of the 6N7 instead of taking its normal course through the audio volume control, and thus the noise pulse is shorted out.

The circuit is completely self-adjusting as to received carrier strength, and gives equal suppression regardless of the carrier level.

Series-Valve Limiters. In the *Bacon* series-tube limiter circuits, the normal signal is carried by the cathode-to-plate current of an additional diode connected into the circuit. This cathode-to-plate current can only flow as long as the plate is positive with respect to the cathode of the diode. Hence, by limiting the range of input signal voltages over which this plate current will flow in conformity with the polarity of the noise pulses as they will appear in the output of the detector, noise limiting will be obtained by adjusting the voltages to such a point that all incoming pulses greater than those produced by 100 per cent modulation of the incoming carrier will cause the plate to go negative with respect to the cathode of the noise diode. The strong noise pulses will then find an open circuit in their path from detector to audio amplifier, although noise pulses up to and including the amplitude of the incoming signal (and the incoming signal) will be passed on to the audio stages.

In a conventional diode detector, the noise pulses will be increasingly negative with respect to normal signal levels, so it is necessary to feed the audio into the plate of the limiter diode and to run the cathode of this diode negative with respect to the plate. This arrangement is shown in Figure 36. The amount of bias is adjusted manually so that all normal signal strengths will be handled, but that pulses in excess of this strength will cause the plate to go negative with respect to the cathode and cause the pulse to be limited in amplitude.

In a power detector, or infinite-impedance detector, the noise pulses are positive with respect to normal signals. In this case it is necessary to feed the detector output into the cathode of the diode limiter, and to bias the plate a certain fixed amount positive with respect to the cathode, as shown in Figure 37. Then, with noise pulses which exceed the positive bias, which has been manually adjusted to appear on the plate, the cathode will go positive with respect to the plate, and the continuity of the signal will be stopped.

A disadvantage of all series-tube noise lim-

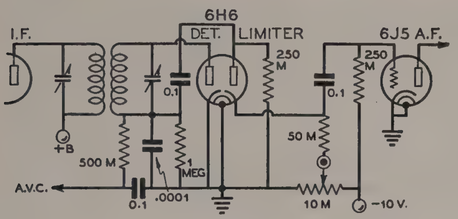


Figure 36.
BACON SERIES LIMITER.
The series type of limiter breaks the circuit between the detector and first audio stage on noise peaks.

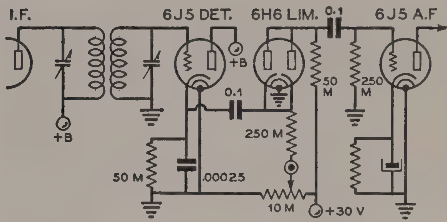


Figure 37.
SERIES LIMITER WITH INFINITE-IMPEDANCE DETECTOR.
This arrangement of the series limiter is applicable to both infinite-impedance and power detectors.

iters is that the signal strength output of the detector is reduced by a considerable amount, often as much as 8 to 10 db, which sometimes requires an additional audio stage or a high-gain stage in place of a low-gain one.

A more detailed and comprehensive discussion of noise balancing and noise limiting systems will be found in the RADIO *Noise Reduction Handbook.*

Receiver Adjustment

The simplest type of regenerative receiver requires little adjustment other than those necessary to insure correct tuning and smooth regeneration over some desired range. Receivers of the tuned radio-frequency type and superheterodynes require precise alignment to obtain the highest possible degree of selectivity and sensitivity.

Good results can only be obtained from a receiver when it is properly aligned and adjusted. The most practical technique for making these adjustments is given in the following discussion.

Instruments. A very small number of instruments will suffice to check and align any multitube receiver, the most important of these testing units being a modulated oscillator and a d.c. and a.c. voltmeter. The meters are essential in checking the voltage applied at *each* circuit point from the power supply. If the a.c. voltmeter is of the oxide-rectifier type, it can be used, in addition, as an output meter when connected across the receiver output when tuning to a modulated signal. If the signal is a steady tone, such as from a test oscillator, the output meter will indicate the value of the detected signal. In this manner, lineup adjustments may be visually noted on the meter rather than by increases or decreases of sound intensity as detected by ear.

T.R.F. Receiver Alignment. The alignment procedure in a multi-stage t.r.f. receiver is exactly the same as aligning a single stage. If the detector is regenerative, each preceding stage is successively aligned while keeping the detector circuit tuned to the test signal, the latter being a station signal or one locally generated by a test oscillator loosely coupled to the antenna lead. During these adjustments, the r.f. amplifier gain control is adjusted for maximum sensitivity, assuming that the r.f. amplifier is stable and does not oscillate. Oscillation is indicative of improper by-passing or shielding. Often a sensitive receiver can be roughly aligned by tuning for maximum noise pickup, such as parasitic oscillations originating from static or electrical machinery.

Superheterodyne Alignment. Aligning a superhet is a detailed task requiring a great amount of care and patience. It should never be undertaken without a thorough understanding of the involved job to be done and then only when there is abundant time to devote to the operation. There are no short cuts; every circuit must be adjusted individually and accurately if the receiver is to give peak performance. The precision of each adjustment is dependent upon the accuracy with which the preceding one was made.

Superhet alignment requires (1) a good signal generator (modulated oscillator) covering the radio and intermediate frequencies and equipped with an attenuator and B-plus switch; (2) the necessary socket wrenches, screwdrivers, or "neutralizing tools" to adjust the various i.f. and r.f. trimmer condensers; and (3) some convenient type of tuning indicator, such as a copper-oxide or electronic voltmeter.

Throughout the alignment process, unless specifically stated otherwise, the a.f. and r.f. gain controls must be set for maximum output, the beat oscillator switched off, the signal-strength meter cut out, the crystal filter set for minimum selectivity, and the a.v.c. turned off. If no provision is made for a.v.c. switching, the signal generator output must be reduced to a low level by means of the attenuator. When the signal output of the receiver is ex-

cessive, either the attenuator or the a.f. gain control may be turned down, but never the r.f. gain control.

I.F. Alignment. After the receiver has been given a rigid electrical and mechanical inspection, and any faults which may have been found in wiring or the selection and assembly of parts corrected, the i.f. amplifier may be aligned as the first step in the checking operations.

The coils for the r.f. (if any), mixer, and high-frequency oscillator stages must be in place. It is immaterial which coils are inserted, since they will serve during the i.f. alignment only to prevent open-grid hum pick-up or oscillation.

With the signal generator set to give a modulated signal on the frequency at which the i.f. amplifier is to operate, clip the output leads from the generator to the last i.f. stage—"hot" end through a small fixed condenser to the control grid, "cold" end to the receiver ground. Adjust both trimmer condensers in the last i.f. transformer (the one between the last i.f. transformer and the second detector) to resonance as indicated by signal peak in the headphones or speaker and maximum deflection of the output meter.

Each i.f. stage is adjusted in the same manner, moving the hot lead, stage by stage, back toward the front end of the receiver and backing off the attenuator as the signal strength increases in each new position. The last adjustment will be made to the first i.f. transformer with the hot signal generator lead connected to the control grid of the mixer. Occasionally it is necessary to disconnect the mixer grid lead from the coil, grounding it through a 1,000- or 5,000-ohm resistor, and coupling the signal generator through a small capacitance to the grid.

When the last i.f. adjustment has been completed, it is good practice to go back through the i.f. channel, re-peaking all of the transformers. It is imperative that this recheck be made in sets which do not include a crystal filter, and where necessarily the simple alignment of the i.f. amplifier to the generator is final.

I.F. with Crystal Filter. There are several ways of aligning an i.f. channel which contains a crystal-filter circuit. However, the following method is one which has been found to give satisfactory results in every case: if the i.f. channel is known to be far out of alignment, or if the initial alignment of a new receiver is being attempted, the crystal itself should first be used to control the frequency of a test oscillator. The circuit shown in Figure 38 can be used. A b.f.o. coil, as shown in the diagram, can be used for the plate inductance. If none is handy, one winding of an i.f. transformer may be used. In either case, the trimmer across the winding may usually be used as the tuning condenser indicated in the diagram.

A milliammeter inserted in the plate circuit will indicate oscillation, the plate current dipping as the condenser tunes the inductance to the resonant frequency of the crystal. Some crystals will require additional grid-plate capacity for oscillation; if so, a 30-$\mu\mu$fd. mica trimmer may be connected from plate to grid of the oscillator tube. The oscillator is then used as a line-up oscillator, as described in the preceding section, by using a.c. for plate supply instead of batteries. The a.c. plate supply gives a modulated signal suitable for the preliminary lining-up process.

For the final i.f. alignment, the crystal should be replaced in the receiver and the phasing condenser set at the "phased" setting, if this is known. If the proper setting of the phasing condenser is unknown, it can be set at half capacity to start with. Next, the output from a signal generator should be applied between the mixer grid and ground, and, with the receiver's a.v.c. circuit operating and the beat oscillator turned "off," the signal generator should be slowly tuned across the i.f. amplifier frequency.

As the generator is tuned through the crystal frequency, the receiver's signal strength meter will give a sudden kick. Should the receiver not be provided with a signal-strength meter, a vacuum-tube voltmeter, such as shown in Chapter 24, can be connected across the a.v.c. line; if the receiver has neither a.v.c. nor a tuning meter, the vacuum-tube voltmeter may be connected between the second detector grid and ground. In any case, a kick of either the tuning meter or the vacuum-tube voltmeter will indicate crystal resonance. It is quite probable that more than one resonance point will be found if the receiver is far out of alignment. The additional points of resonance are spurious crystal peaks; the strongest peak should be chosen, and the signal generator left tuned to this frequency.

The phasing condenser should next be adjusted for *minimum* hiss or noise in the receiver output, and the selectivity control, if any is provided, set for maximum selectivity. The signal generator should now be carefully tuned to the exact peak of the crystal. From this point on, the alignment of the i.f. amplifier follows conventional practice, except that the a.v.c. circuit is used as an alignment indicator, each circuit being adjusted for maximum output. If the receiver is of the type having no a.v.c. or tuning indicator, and the vacuum-tube voltmeter must be connected

across the second-detector grid circuit, it will be necessary to remove the vacuum-tube voltmeter and make the final adjustment on the last i.f. transformer by ear, after the other transformers have been aligned.

B. F. O. Adjustment. Adjusting the beat oscillator is relatively simple. It is only necessary to tune the receiver to resonance with any signal, as indicated by the tuning indicator, and then turn on the b.f.o. and set its trimmer (or trimmers) to produce the desired beat note. Setting the beat oscillator in this way will result in the beat note being stronger on one "side" of the signal than on the other, which is what is desired for maximum selectivity. The b.f.o. should *not* be set to "zero beat" with the receiver tuned to resonance with the signal, as this will cause an equally strong beat to be obtained on both sides of resonance.

Front-End Alignment. The alignment of the "front end" of a manufactured receiver is a somewhat involved process, and varies considerably from one receiver to another and for that reason will not be discussed here. Those interested in the alignment of such receivers usually will find full instructions in the operating manual or instruction book supplied with the receiver. Likewise, full alignment data are always given when an "all wave" tuning assembly for incorporation in home-built receivers is purchased.

In aligning the front end of a home-constructed superheterodyne which covers only the amateur bands, the principal problems are those of securing proper bandspread in the oscillator, and then tracking the signal-frequency circuits with the oscillator. The simplest method of adjusting the oscillator for proper bandspread is to tune in the oscillator on an "all wave" receiver, and adjust its bandspread so that it covers a frequency range equal to that of the tuning range desired in the receiver but over a range of frequencies equal to the desired signal range plus the intermediate frequency. For example: if the receiver is to tune from 13,950 to 14,450 kc. to cover the 14-Mc. amateur band with a 50-kc. leeway at each end, and the intermediate frequency is 465 kc., the oscillator should tune from 13,950+465 kc. to 14,450+465 kc., or from 14,415 to 14,915 kc.

(Note: The foregoing assumes that the oscillator will be operated on the high-frequency side of the signal, which is the usual condition. It is quite possible, however, to have the oscillator on the low-frequency side of the signal, and if this is desired, the intermediate frequency is simply *subtracted* from the signal frequency, rather than added, to give the required oscillator frequency).

If no calibrated auxiliary receiver is available, the following procedure should be used to adjust the oscillator to its proper tuning range: a modulated signal from the signal generator is fed into the mixer grid, with mixer grid coil for the band being used in place, and with the signal generator set for the highest frequency in the desired tuning range and the bandspread condenser in the receiver set at minimum capacity. Next, the oscillator bandsetting condenser is slowly decreased from maximum capacity until a strong signal from the signal generator is picked up. The first strong signal picked up will be when the oscillator is on the low-frequency side of the signal. If it is desired to use this beat, the oscillator bandsetting condenser need not be adjusted further. However, if it is intended to operate the oscillator on the high-frequency

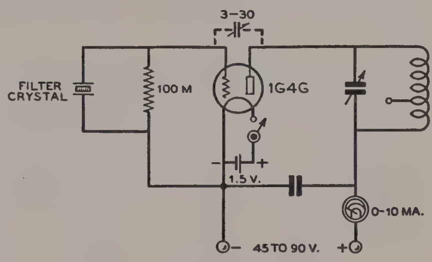


Figure 38.
CRYSTAL TEST OSCILLATOR CIRCUIT.
The receiver's crystal may be placed in this oscillator for a rough alignment of the i.f. amplifier to the crystal frequency. The tank circuit may be a b.f.o. transformer or one winding of an i.f. transformer.

side of the signal, in accordance with usual practice, the bandsetting condenser should be decreased in capacity until the second strong signal is heard. When the signal is properly located, the mixer grid should be next tuned to resonance by adjusting its padder condenser for maximum signal strength.

After the high-frequency end of the band has thus been located, the receiver bandspread condenser should be set at maximum capacity, and the signal generator slowly tuned toward the low-frequency end of its range until its signal is again picked up. If the bandspread adjustment happens to be correctly made, the signal generator calibration will show that it is at the low-frequency end of the desired tuning range. If calibration shows that the low-frequency end of the tuning range falls either higher or lower than what is desired, it will be

necessary to make the required changes in the bandspread circuit described in the preceding section on *Bandspread,* and repeat the checking process until the tuning range is correct.

Tracking. After the oscillator has been set so that it covers the correct range, the tracking of the mixer tuning may be tackled. With the signal generator set to the high-frequency end of the tuning range and *loosely* coupled to the mixer grid, the signal from the generator should be tuned in on the receiver, and the mixer padding condenser adjusted for maximum output. Next, both the receiver and the signal generator should be tuned to the low-frequency end of the receiver's range, and a check made to see if it is necessary to reset the mixer padder to secure maximum output. If the tracking is correct, it will be found that no change in the padder capacity will be necessary. If, however, it is found that the output may be increased by retuning the padder, it will be necessary to readjust the mixer bandspread.

An increase in signal strength with an increase in padding capacity indicates that the bandspread is too great, and it will be necessary to increase the tuning range of the mixer. An increase in signal strength with a decrease in padding capacity shows that the mixer tuning range is too great, and the bandspread will have to be increased.

When the mixer bandspread has been adjusted so that the tracking is correct at both ends of a range as narrow as an amateur band, it may be assumed that the tracking is nearly correct over the whole band. The signal generator should then be transferred to the grid of the r.f. stage, if the receiver has one, and the procedure described for tracking the mixer carried out in the r.f. stage.

Series Tracking Condensers. The above discussion applies solely to receivers in which a small tuning range is covered with each set of coils, and where the ranges covered by the oscillator and mixer circuits represent nearly equal percentages of their operating frequencies, i.e., the intermediate frequency is low. When these conditions are not satisfied, such as in continuous-coverage receivers and in receivers in which the intermediate frequency is a large proportion of the signal frequency, it becomes necessary to make special provisions for oscillator tracking. These provisions usually consist of ganged tuning condensers in which the oscillator section plates are shaped differently and have a different capacity range than those used across the other tuned circuits, or the addition of a "tracking condenser" in series with the oscillator tuning condenser in conjunction with a smaller coil.

While series tracking condensers are seldom used in home-constructed receivers, it may sometimes be necessary to employ one, as in, for example, a receiver using a 1600-kc. i.f. channel and covering the 3500–4000 kc. amateur band. The purpose of the series tracking condenser is to slow down the oscillator's tuning rate when it operates on the high-frequency side of the signal. This method allows perfect tracking at three points throughout the tuning range. The three points usually chosen for the perfect tracking are at the two ends and center of the tuning range; between these points the tracking will be close enough for all practical purposes.

In home-constructed sets, the adjustment of the tracking condenser and oscillator coil inductance is largely a matter of cut-and-try, requiring a large amount of patience and an understanding of the results to be expected when the series capacity and the oscillator inductance are changed.

Receivers with A.V.C. When lining up a receiver which has automatic volume control (a.v.c.) which cannot be turned off, it is considered good practice to keep the test oscillator signal near the threshold sensitivity at all times to give the effect of a very weak signal, thus making it possible to detect changes in the receiver output level during alignment.

Testing. In checking over a receiver, certain troubles are often difficult to locate. By making voltage or continuity tests, blown-out condensers, or burned-out resistors, coils, or transformers may usually be located. Oscillators are usually checked by means of a d.c. voltmeter connected from ground to screen or plate-return circuits. Short-circuiting the tuning condenser plates usually should produce a change in voltmeter reading. A vacuum-tube-type voltmeter is very handy for the purpose of measuring the correct amount of oscillator r.f. voltage supplied to the first detector circuit. The proper value of the r.f. voltage is approximately 1 volt less than the fixed grid bias on the first detector when the voltage is introduced into either the grid or the cathode circuit.

Incorrect voltages, poor resistors, or leaky by-pass or blocking condensers will ruin the audio tone of the receiver. Defective tubes can be checked in a tube tester. Loud-speaker rattle is not always a defect in the voice coil or spider support, or metallic filings in its air gap; more often the distortion is caused by overloading the audio amplifier. An i.f. amplifier can also impair splendid tone due to a defective tube or overloading.

It is a good idea to have all tubes in a receiver checked periodically, because if a tube slowly becomes noisy, soft, or deficient in emission, the operator may not realize that the

performance is not up to the full capabilities of the receiver. Any tube which does not test up to the equivalent of a new tube should be replaced, as a tube that once starts to "go" cannot possibly give very many more hours of useful service.

On the other hand, there is little point in replacing all tubes periodically, because tests have shown that a tube that has been in use for three or four years, if it still is giving satisfactory service, is just as likely to provide another year of uninterrupted service as is a brand new tube.

It should be borne in mind that electrolytic condensers, even of the best quality, have a limited life—the length of useful service depending upon the quality and application of the condenser. Unlike tubes, electrolytic condensers seldom give any trouble in the first three years of use (if of good quality and not overloaded). However, they seldom last more than five years, unless they are the less commonly used "wet" type. For this reason it is advisable to replace all electrolytic condensers in a receiver every four or five years if reliability of service is important.

CHAPTER FIVE

Radio Receiving Tube Characteristics

FOOTNOTE references for both standard and special receiving tubes will be found immediately following the socket connection diagrams for these tubes. Footnote references for various cathode-ray tubes will be found immediately following the separate group of socket connections for cathode-ray tubes.

A suffix (G) in parentheses after a standard octal base tube indicates that the tube also is manufactured with glass envelope, a suffix (GT) indicating that the tube also is manufactured with small tubular glass envelope. Thus 6J5 (G) (GT) indicates that this tube is available with metal, glass, or small tubular glass envelope; 6AG7 indicates that this tube is available only in metal; and 5Y3-G indicates that this tube is available only in glass.

The "Bantam" line of GT type tubes by one manufacturer have a metal shell base which is connected to the pin which would ground the shell of an equivalent metal tube. A sleeve shield slipped over the tube thus is automatically grounded.

Several manufacturers supply certain of their tubes with ceramic base at a slight increase in the price. The ceramic base ordinarily is indicated by the presence of the letter "X" at the end of the regular type number.

Certain of the "7" series of tubes have a nominal heater rating of 7 volts instead of the usual 6.3-volt rating. The heater is the same, however, and either the "6" series or the "7" series may be used on either 6.3 or 7 volts. To simplify the tables, all such tubes are shown with a rating of 6.3 volts. The same applies to certain of the "14" series of tubes, these tubes having the same heater as corresponding tubes of the "12" series but a nominal heater rating of 14 volts instead of 12.6 volts.

Socket terminals shown as unused in the table of socket connections should not be used as tie-points for other wiring unless the tube has no corresponding pin, because "dead" pins are sometimes used as element supports.

When a "G" or "GT" octal base tube is used, the shell grounding terminal (usually pin no. 1) for the corresponding metal counterpart should be connected to ground the same as for a metal tube, as many "G" and "GT" types contain an internal shield.

Tube Base Connections

There are from 4 to 8 pins on tube bases. With the exception of the 5- and 8-prong types of bases the filament or heater pins are those which are heavier than the others.

With the exception of the octal (8-pin) base, the numbering system for the pins is as follows (viewing the tube or socket from the bottom, and with the two heavier heater [or filament] pins horizontal): the no. 1 pin is the left-hand heater or filament pin. Pins no. 2, 3, and so forth follow around in a clockwise direction, the highest number being the right-hand cathode pin. Octal (8-pin) numbers start with no. 1 which is the first pin to the left of the key.

The letters F-F or H-H designate filament or heater, C or K for the cathode, P for the plate, etc., in socket connection or wiring diagrams. The grids of multigrid tubes are numbered with respect to the position they occupy: no. 1 grid is closest to the cathode, no. 2 next closest, etc. When it is desirable that certain elements have a very low capacity with respect to other elements within the tube, they are sometimes terminated in a lead brought out to a cap on top of the tube.

This chapter includes data on receiving tube characteristics, tube socket connections, special purpose and cathode-ray tubes, and cathode-ray socket connections.

TYPE	DESIGN	SOCKET CONN.	CATHODE TYPE AND RATING C.T.	VOLTS	AMP	USED AS	PLATE SUPPLY VOLTS	GRID BIAS (2) VOLTS	SCREEN SUPPLY VOLTS	SCREEN CURRENT MILLAMP.	PLATE CURRENT MILLIAMP.	A.C. PLATE RESISTANCE OHMS	TRANSCONDUCTANCE (GRID-PLATE) µMHOS	AMPLIFICATION FACTOR	LOAD FOR STATED POWER OUTPUT OHMS	POWER OUTPUT WATTS
00-A	DETECTOR TRIODE	4D	D.C. F	5.0	0.25	GRID-LEAK DETECTOR	45	GRID RETURN TO (−) FILAMENT			1.5	30000	666	20	—	—
01-A	DETECTOR AMPLIFIER	4D	D.C. F	5.0	0.25	CLASS A AMPLIFIER	90 135	−4.5 −9.0	—	—	2.5 3.0	11000 10000	725 800	8.0 8.0	—	—
0A4-G	GAS-TRIODE	4V	COLD	—	—	RELAY SERVICE	PEAK CATHODE CURRENT, 100 MAX. MA. D.C. CATHODE CURRENT, 25 MAX. MA. STARTER-ANODE DROP, 60 APPROX. VOLTS. ANODE DROP, 70 APPROX. VOLTS.									
0Z3	FULL-WAVE GAS RECTIFIER	5N	COLD	—	—	RECTIFIER	FOR OTHER CHARACTERISTICS REFER TO TYPE 0Z4									
0Z4 (G)	FULL WAVE GAS RECTIFIER	4R	COLD	—	—	RECTIFIER	STARTING-SUPPLY VOLTAGE PER PLATE, 300 MIN. PEAK VOLTS. PEAK PLATE CURRENT, 200 MAX. MA. D.C. OUTPUT CURRENT, 75 MAX., 30 MIN. MA. D.C. OUTPUT VOLTAGE, 300 MAX. VOLTS.									
1A4-P	SUPER-CONTROL R.F. AMPLIFIER PENTODE	4M	D.C. F	2.0	0.06	AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 1D5-GP									
1A4-T	SUPER-CONTROL R.F. AMPLIFIER TETRODE	4K	D.C. F	2.0	0.06	AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 1D5-GT									
1A5-GT/G	POWER AMPLIFIER PENTODE	6X	D.C. F	1.4	0.05	CLASS A AMPLIFIER	85 90	−4.5 −4.5	85 90	0.7 0.8	3.5 4.0	300000 300000	800 850	—	25000 25000	0.100 0.115
1A6	PENTAGRID CONVERTER (8)	6L	D.C. F	2.0	0.06	CONVERTER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 1D7-G.									
1A7-G (GT)	PENTAGRID CONVERTER (8)	7Z	D.C. F	1.4	0.05	CONVERTER	90	0	45 (14)	0.6	0.55	600000	ANODE-GRID (#2): 90 MAX. VOLTS, 1.2 MA. OSCILLATOR-GRID (#1) RESISTOR, 0.2 MEG. CONVERSION TRANSCOND., 250 MICROMHOS.			
1B4-P	R.F. AMPLIFIER PENTODE	4M	D.C. F	2.0	0.06	AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 1E5-GP.									
1B5/25S	DUPLEX-DIODE TRIODE	6M	D.C. F	2.0	0.06	TRIODE UNIT AS AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 1H6-G.									
1B7-G	PENTAGRID CONVERTER	7Z	D.C. F	1.4	0.10	OSCILLATOR-AMPLIFIER CONVERTER	90	GRID RETURNS THRU 200000 Ω. RESISTOR TO (−F)	45	1.3	1.5	350000	350	GRID N°2, 90 VOLTS, 1.6 MA.		
1B8-GT	MULTI-PURPOSE	8AW	D.C. F	1.4	0.10	DIODE-TRIODE BEAM AMPLIFIER	90 90	TRIODE, 0 BEAM AMP., −6	90	1.4	0.15 6.3	240000	275 1150	—	14000	210 MW.
1C4	SUPER-CONTROL R.F. AMPLIFIER PENTODE	4M	D.C. F	2.0	0.12	AMPLIFIER	180	0	67.5	0.9	2.5	1000000	1000	1000	—	—
1C5-GT/G	POWER AMPLIFIER PENTODE	6X	D.C. F	1.4	0.10	CLASS A AMPLIFIER	83 90	−7.0 −7.5	83 90	1.6 1.6	7.0 7.5	110000 115000	1500 1550	—	9000 8000	0.20 0.24
1C6	PENTAGRID CONVERTER (8)	6L	D.C. F	2.0	0.12	CONVERTER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 1C7-G.									
1C7-G	PENTAGRID CONVERTER (8)	7Z	D.C. F	2.0	0.12	CONVERTER	135 180	−3.0 −3.0	67.5 67.5	2.5 2.0	1.3 1.5	600000 700000	ANODE GRID (#2): 180 (3) MAX. VOLTS, 4.0 MA. OSCILLATOR-GRID (#1) RESISTOR. (19) CONVERSION TRANSCOND., 325 MICROMHOS.			
1D4	POWER AMPLIFIER PENTODE	5B	D.C. F	2.0	0.24	CLASS A AMPLIFIER	180	−6.0	180	2.3	9.5	137000	—	330	15000	0.75
1D5-GP	SUPER-CONTROL R.F. AMPLIFIER PENTODE	5Y	D.C. F	2.0	0.06	CLASS A AMPLIFIER	90 180	{ −3.0 MIN. }	67.5 67.5	0.9 0.8	2.2 2.3	600000 1000000	720 750	425 750	—	—
1D5-GT	SUPER CONTROL R.F. AMPLIFIER TETRODE	5R	D.C. F	2.0	0.06	AMPLIFIER	135 180	−3.0 −3.0	67.5 67.5	0.7 0.7	2.2 2.2	350000 600000	625 659	—	—	—

1D7-G	PENTAGRID CONVERTER (8)	7Z	D.C. F	2.0	0.06	CONVERTER	135 180	{-3.0 MIN.}	67.5 67.5	2.5 2.4	1.2 1.3	400000 500000	ANODE-GRID (#2): 180 (3) MAX. VOLTS. 2.3 MA. OSCILLATOR-GRID (#1) RESISTOR. (19) CONVERSION TRANSCOND., 300 MICROMHOS.			
1D8-GT	DIODE-TRIODE POWER AMPLIFIER PENTODE	8AJ	D.C. F	1.4	0.1	PENTODE UNIT AS CLASS A AMPLIFIER	45 90	-4.5 -9.0	45 90	0.3 1.0	1.6 5.0	300000 200000	650 925	——	20000 12000	0.035 0.200
						TRIODE UNIT AS CLASS A AMPLIFIER	45 90	0 0	——	——	0.3 1.1	77000 43500	325 575	25 25	——	——
1E4-G	GENERAL PURPOSE TRIODE	5S	D.C. F	1.4	0.05	AMPLIFIER	90	-3.0	——	——	1.5	17000	825	14	——	——
1E5-GP	R.F. AMPLIFIER PENTODE	5Y	D.C. F	2.0	0.06	CLASS A AMPLIFIER	90 180	-3.0 -3.0	67.5 67.5	0.7 0.6	1.6 1.7	1000000 1500000	600 650	550 1000	——	——
1E7-G	TWIN-PENTODE POWER AMPLIFIER	8C	D.C. F	2.0	0.24	PUSH-PULL CLASS A AMPLIFIER	135	-7.5	135	2.0	7.0	260000	1425	——	24000	0.575 (36)
1F4	POWER AMPLIFIER PENTODE	5K	D.C. F	2.0	0.12	AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 1F5-G.									
1F5-G	POWER AMPLIFIER PENTODE	6X	D.C. F	2.0	0.12	CLASS A AMPLIFIER	90 135	-3.0 -4.5	90 135	1.1 2.4	4.0 8.0	240000 200000	1400 1700	——	20000 16000	0.11 0.31
1F6	DUPLEX-DIODE PENTODE	6W	D.C. F	2.0	0.06	PENTODE UNIT AS AMPLIFIER		FOR OTHER CHARACTERISTICS, REFER TO TYPE 1F7-GV.								
1F7-G (GV)	DUPLEX-DIODE PENTODE	7AD	D.C. F	2.0	0.06	PENTODE UNIT AS R.F. AMPLIFIER	180	-1.5	67.5	0.7	2.2	1000000	650	——	——	——
						PENTODE UNIT AS A.F. AMPLIFIER	135 (18)	-2.0	SCREEN SUPPLY, 135 VOLTS APPLIED THROUGH 0.8-MEGOHM RESISTOR. GRID RESISTOR (10) 1.0 MEGOHM. VOLTAGE GAIN, 46.							
1G4-GT/G	DETECTOR AMPLIFIER TRIODE	5S	D.C. F	1.4	0.05	CLASS A AMPLIFIER	90	-6.0	——	——	2.3	10700	825	8.8	——	——
1G5-G	POWER AMPLIFIER PENTODE	6X	D.C F	2.0	0.12	CLASS A AMPLIFIER	90 135	-6.0 -13.5	90 135	2.5 2.5	8.5 8.7	133000 160000	1500 1550	——	8500 9000	0.25 0.55
1G6-GT/G	TWIN-TRIODE AMPLIFIER	7AB	D.C. F	1.4	0.10	CLASS B AMPLIFIER	90	0	——	——	2.0	——	——	——	12000	0.675 (36)
1H4-G	DETECTOR AMPLIFIER TRIODE (1)	5S	D.C. F	2.0	0.06	CLASS A AMPLIFIER	90 135 180	-4.5 -9.0 -13.5	——	——	2.5 3.0 3.1	11000 10300 10300	850 900 900	9.3 9.3 9.3	——	——
						CLASS B AMPLIFIER	157.5	-15.0	——	——	1.0 (13)	——	——	——	8000	2.1 (12)
1H5-G (GT)	DIODE HIGH-MU TRIODE	5Z	D.C. F	1.4	0.05	TRIODE UNIT AS CLASS A AMPLIFIER	90	0	——	——	0.15	240000	275	65	——	——
1H6-G	DUPLEX-DIODE TRIODE	7AA	D.C. F	2.0	0.06	TRIODE UNIT AS CLASS A AMPLIFIER	135	-3.0	——	——	0.8	35000	575	20	——	——
1J5-G	POWER AMPLIFIER PENTODE	6X	D.C. F	2.0	0.12	CLASS A AMPLIFIER	135	-16.5	135	1.8	7.0	125000	1000	125	13500	0.450
1J6-G	TWIN-TRIODE AMPLIFIER	7AB	D.C. F	2.0	0.24	CLASS B AMPLIFIER	135 135	0 -3.0	——	——	(37) 5.0	——	——	——	10000 10000	2.1 1.9 (36)
1LA4	POWER AMPLIFIER PENTODE	5AD	D.C. F	1.4	0.05	CLASS A AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 1A5-GT/1A5-G									
1LA6	PENTAGRID CONVERTER	7AK	D.C. F	1.4	0.05	CONVERTER	90	0	45	0.6	0.55	750000	250	OSC. GRID THRU 200000 Ω TO -F ANODE GRID 90 V., 1.2 MA.		
1LB4 (G)	POWER AMPLIFIER PENTODE	5AD	D.C. F	1.4	0.05	CLASS A AMPLIFIER	90	-9.0	90	1.0	5.0	200000	925	——	12000	0.20
1LB6-GL	PENTAGRID CONVERTER	8AX	D.C. F	1.4	0.05	CONVERTER	90	0	67.5	2.2	0.4	2000000	100	ANODE GRID 67.5 V., 1.2 MA.		
1LC5	R.F. AMPLIFIER PENTODE	7AO	D.C. F	1.4	0.5	AMPLIFIER	90	0	45	0.2	1.15	1500000	775	——	——	——
1LC6	PENTAGRID CONVERTER (8)	7AK	D.C. F	1.4	0.05	CONVERTER	45	0	35	1.4	0.7	300000	ANODE-GRID (#2): 45 MAX. VOLTS, 1.4 MA. OSCILLATOR-GRID (#1) RESISTOR, 0.2 MEG. CONVERSION TRANSCOND., 250 MICROMHOS.			
1LD5	DIODE-PENTODE	6AX	D.C. F	1.4	0.05	PENTODE UNIT AS AMPLIFIER	90	0	45	0.1	0.6	750000	575	——	——	——
1LE3-GL	TRIODE	4AA	FOR OTHER CHARACTERISTICS, REFER TO TYPE 1E4-G													
1LH4	DIODE HIGH-MU TRIODE	5AG	D.C. F	1.4	0.05	TRIODE UNIT AS CLASS A AMPLIFIER	90	0	——	——	0.15	240000	275	65	——	——
1LN5	R.F. AMPLIFIER PENTODE	7AO	D.C. F	1.4	0.05	FOR OTHER CHARACTERISTICS, REFER TO TYPE 1N5-G										

TYPE	DESIGN	SOCKET	CATHODE TYPE AND RATING			USED AS	PLATE SUPPLY VOLTS	GRID BIAS (2) VOLTS	SCREEN SUPPLY VOLTS	SCREEN CURRENT MILLIAMP.	PLATE CURRENT MILLIAMP.	A.C. PLATE RESISTANCE OHMS	TRANSCONDUCTANCE (GRID-PLATE) µMHOS	AMPLIFICATION FACTOR	LOAD FOR STATED POWER OUTPUT OHMS	POWER OUTPUT WATTS
			C.T.	VOLTS	AMP.											
1N5-G (GT)	R.F. AMPLIFIER PENTODE	5Y	D.C. F	1.4	0.05	CLASS A AMPLIFIER	90	0	90	0.3	1.2	1500000	750	—	—	—
1N6-G (GT)	DIODE-POWER AMPLIFIER-PENTODE	7AM	D.C. F	1.4	0.05	PENTODE UNIT AS A.F. AMPLIFIER	90	−4.5	90	0.6	3.1	300000	800	—	25000	0.10
1P5-G (GT)	R.F. AMPLIFIER PENTODE	5Y	D.C. F	1.4	0.05	CLASS A AMPLIFIER	90	0	90	0.7	2.3	800000	750	640	—	—
1Q5-GT/G	BEAM POWER AMPLIFIER	6AF	D.C. F	1.4	0.10	CLASS A AMPLIFIER	90	−4.5	90	1.6	9.5	—	2100	—	8000	0.27
1R5	PENTAGRID CONVERTER	7AT	D.C. F	1.4	0.05	CONVERTER	45 90	0 0	45 45	1.9 1.8	0.7 0.8	600000 750000	GRID #1 RESISTOR, 100 000 OHMS. CONVERSION TRANSCOND., 250 µMHOS			—
1S4	POWER AMPLIFIER PENTODE	7AV	D.C. F	1.4	0.10	CLASS A AMPLIFIER	45	−4.5	45	0.8	3.8	100000	1250	—	8000	0.065
1S5	DIODE-PENTODE	6AU	D.C. F	1.4	0.05	PENTODE UNIT AS CLASS A AMPLIFIER	PLATE SUPPLY, 41 VOLTS APPLIED THROUGH 1 MEGOHM RESISTOR. SCREEN SUPPLY, 41 VOLTS GRID BIAS, 0 VOLTS. GRID RESISTOR, 10 MEGOHMS. VOLTAGE GAIN, 30 APPROXIMATELY. (17)									
1SA6-GT	R.F. AMPLIFIER PENTODE	6BD	D.C. F	1.4	0.05	AMPLIFIER	90	0	67.5	0.68	2.45	800000	970	—	—	—
1SB6-GT	DIODE-PENTODE	6BE	D.C. F	1.4	0.05	PENTODE UNIT AS AMPLIFIER	90	0	67.5	0.38	1.45	700000	665	—	—	—
1T4	SUPER-CONTROL R.F. AMPLIFIER PENTODE	6AR	D.C. F	1.4	0.05	CLASS A AMPLIFIER	45 90	0 0	45 45	0.7 0.65	1.9 2.0	350000 800000	700 750	—	—	—
1T5-GT	BEAM POWER AMPLIFIER	6X	D.C. F	1.4	0.05	CLASS A AMPLIFIER	90	−6.0	90	1.4	6.5	—	1150	—	14000	0.17
1-v	HALF-WAVE RECTIFIER	4G	H	6.3	0.3	WITH CONDENSER-INPUT FILTER	MAX. A.C. PLATE VOLTS (RMS) 325 MAX. D.C. OUTPUT MA., 45				MINIMUM TOTAL EFFECTIVE PLATE-SUPPLY IMPEDANCE: UP TO 117 VOLTS, 0 OHMS; AT 150 VOLTS, 30 OHMS; AT 325 VOLTS, 75 OHMS					
2A3	POWER AMPLIFIER TRIODE	4D	F	2.5	2.5	CLASS A AMPLIFIER	250	−45.0	—	—	60.0	800	5250	4.2	2500	3.5
						PUSH-PULL CLASS AB_1 AMPLIFIER	300 300	CATHODE BIAS, 780 OHMS (13) −62 VOLTS, FIXED BIAS			80.0 (13) 80.0 (13)	—	—	—	5000 3000	10.0 (12) 15.0 (12)
2A5	POWER AMPLIFIER PENTODE	6B	H	2.5	1.75	AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6F6.									
2A6	DUPLEX-DIODE HIGH-MU TRIODE	6G	H	2.5	0.8	TRIODE UNIT AS AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6SQ7									
2A7 (S)	PENTAGRID CONVERTER (8)	7C	H	2.5	0.8	CONVERTER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6A8.									
2B7	DUPLEX-DIODE PENTODE	7D	H	2.5	0.8	PENTODE UNIT AS AMPLIFIER	FOR OTHER CHARACTERISTICS REFER TO TYPE 6B8-G									
2E5	ELECTRON-RAY TUBE	6R	H	2.5	0.8	TUNING INDICATOR	FOR OTHER CHARACTERISTICS REFER TO TYPE 6E5.									
2G5	ELECTRON-RAY TUBE	6R	H	2.5	0.8	VISUAL INDICATOR	250	−22 FOR 0° SHADOW ANGLE	TARGET 250	PLATE VOLTAGE SUPPLIED THROUGH 1 MEG. RESISTOR						
2W3 (GT)	HALF-WAVE RECTIFIER	4X	F	2.5	1.5		MAXIMUM A.C. VOLTAGE .... 350 VOLTS, RMS MAXIMUM D.C. OUTPUT CURRENT .... 55 MILLIAMPERES									
2Y2	HALF-WAVE RECTIFIER	4AB	H	2.5	1.75		MAXIMUM A.C. VOLTAGE .... 4500 VOLTS, RMS MAXIMUM D.C. OUTPUT CURRENT .. .... 5.0 MILLIAMPERES									

3A8-GT	DIODE-TRIODE R F. PENTODE	8AS	D.C. F	1.4	0.1	TRIODE UNIT AS CLASS A AMPLIFIER	90	0	———	———	0.2	200000	325	65	———	———
				2.8	0.05	PENTODE UNIT AS CLASS A AMPLIFIER	90	0	90	0.5	1.5	800000	750	———	———	———
3B5-GT	BEAM POWER AMPLIFIER	7AP	D.C. F	1.4 2.8	0.10 0.05	CLASS A AMPLIFIER	67.5	−7.0	67.5	0.5	6.7	100000	1500	———	5000	0.18
3C5-GT	POWER AMPLIFIER PENTODE	7AD	D.C. F	1.4 2.8	0.10 0.05	POWER AMPLIFIER PENTODE	90	−9.0	———	———	6.0	———	1450	———	10000	0.26
3LE4	POWER AMPLIFIER PENTODE	6BA	D.C. F	2.8	0.05	CLASS A AMPLIFIER	90	−9.0	90	1.8	9.0	110000	1600	———	6000	0.30
3Q4	POWER AMPLIFIER PENTODE	7BA	D.C. F	1.4 2.8	0.10 0.05	CLASS A AMPLIFIER	90 90	−4.5 −4.5	90 90	2.1 1.7	9.5 7.7	100000 120000	2150 2000	———	10000 10000	0.27 0.24
3Q5-GT	BEAM POWER AMPLIFIER	7AQ	D C F	1.4 2.8	0.10 0.05	CLASS A AMPLIFIER	90 90	−4.5 −4.5	90 90	1 6 1 0	9.5 7 5	100000 110000	2100 1800	———	8000 8000	0.27 0.25
3S4	POWER AMPLIFIER PENTODE	7BA	D.C. F	1.4 2.8	0.10 0.05	CLASS A AMPLIFIER	90 90	−7.0 −7.0	67.5 67.5	1.4 1.1	7 4 6.1	100000 100000	1575 1425	———	8000 8000	0.27 0.235
4A6-G	TWIN-TRIODE POWER AMPLIFIER	8L	D.C. F	2.0 4.0	0.12 0.06	CLASS B AMPLIFIER	90	− 1.5	ZERO-SIGNAL PLATE CURRENT, 1 1 MA. MAX -SIGNAL PLATE CURRENT, 10.8 MA						8000	1.0
5T4	FULL-WAVE RECTIFIER	5T	F	5.0	2.0	WITH CONDENSER INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 450 MAX. PEAK INVERSE VOLTS., 1550				MAX. D.C. OUTPUT MA., 225 MAX. PEAK PLATE MA., 675			MIN. TOTAL EFFECTIVE SUPPLY IMPEDANCE PER PLATE, 150 OHMS		
						WITH CHOKE INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 550 MAX PEAK INVERSE VOLTS, 1550				MAX. D.C. OUTPUT MA., 225 MAX. PEAK PLATE MA., 675			MINIMUM VALUE OF INPUT CHOKE, 3 HENRIES		
5U4-G	FULL-WAVE RECTIFIER	5T	F	5.0	3.0	WITH CONDENSER INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 450 MAX. PEAK INVERSE VOLTS, 1550				MAX. D.C. OUTPUT MA., 225 MAX PEAK PLATE MA, 675			MIN. TOTAL EFFECTIVE SUPPLY IMPEDANCE PER PLATE, 75 OHMS		
						WITH CHOKE INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 550 MAX. PEAK INVERSE VOLTS, 1550				MAX. D.C. OUTPUT MA., 225 MAX. PEAK PLATE MA., 675			MINIMUM VALUE OF INPUT CHOKE, 3 HENRIES		
5V4-G	FULL-WAVE RECTIFIER	5L	H	5.0	2.0	WITH CONDENSER INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 375 MAX. PEAK INVERSE VOLTS, 1400				MAX D.C. OUTPUT MA., 175 MAX. PEAK PLATE MA., 525			MIN TOTAL EFFECTIVE SUPPLY IMPEDANCE PER PLATE, 65 OHMS		
						WITH CHOKE INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 500 MAX. PEAK INVERSE VOLTS, 1400				MAX D.C OUTPUT MA., 175 MAX. PEAK PLATE MA., 525			MINIMUM VALUE OF INPUT CHOKE, 4 HENRIES		
5W4-GT/G	FULL-WAVE RECTIFIER	5T	F	5.0	1 5	WITH CONDENSER-INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 350 MAX. PEAK INVERSE VOLTS, 1400				MAX. D.C. OUTPUT MA., 100 MAX. PEAK PLATE MA., 300			MIN. TOTAL EFFECTIVE SUPPLY IMPEDANCE PER PLATE, 25 OHMS		
						WITH CHOKE INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 500 MAX. INVERSE PEAK VOLTS, 1400				MAX. D.C. OUTPUT MA., 100 MAX. PEAK PLATE MA., 300			MINIMUM VALUE OF INPUT CHOKE, 6 HENRIES		
5X3	FULL-WAVE RECTIFIER	4C	F	5.0	2.0	RECTIFIER	MAX. A.C. VOLTS PER PLATE (RMS), 400 MAX. D C. OUTPUT MA., 110									
5X4-G	FULL-WAVE RECTIFIER	5Q	F	5.0	3.0	FOR OTHER RATINGS, REFER TO TYPE 5U4-G										
5Y3-GT/ 5Y3-G	FULL-WAVE RECTIFIER	5T	F	5.0	2.0	WITH CONDENSER-INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 350 MAX. PEAK INVERSE VOLTS, 1400				MAX. D.C. OUTPUT MA., 125 MAX. PEAK PLATE MA., 375			MIN. TOTAL EFFECTIVE SUPPLY IMPEDANCE PER PLATE, 10 OHMS		
						WITH CHOKE INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 500 MAX. PEAK INVERSE VOLTS, 1400				MAX. D.C. OUTPUT MA., 125 MAX. PEAK PLATE MA., 375			MINIMUM VALUE OF INPUT CHOKE, 5 HENRIES		
5Y4-G	FULL-WAVE RECTIFIER	5Q	F	5.0	2.0	FOR OTHER RATINGS, REFER TO TYPE 5Y3-G										
5Z3	FULL-WAVE RECTIFIER	4C	F	5.0	3.0	FOR OTHER RATINGS, REFER TO TYPE 5U4-G										
5Z4	FULL-WAVE RECTIFIER	5L	H	5.0	2.0	WITH CONDENSER-INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 350 MAX. PEAK INVERSE VOLTS, 1400				MAX. D.C. OUTPUT MA., 125 MAX. PEAK PLATE MA., 375			MIN. TOTAL EFFECTIVE SUPPLY IMPEDANCE PER PLATE, 30 OHMS		
						WITH CHOKE-INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 500 MAX. PEAK INVERSE VOLTS, 1400				MAX. D.C. OUTPUT MA., 125 MAX. PEAK PLATE MA., 375			MINIMUM VALUE OF INPUT CHOKE, 5 HENRIES		
6A3	POWER AMPLIFIER TRIODE	4D	F	6.3	1.0	POWER AMPLIFIER TRIODE	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6B4-G									
6A4/LA	POWER AMPLIFIER PENTODE	5B	F	6.3	0.3	CLASS A AMPLIFIER	100 180	−6.5 −12.0	100 180	1.6 3.9	9.0 22.0	83250 45500	1200 2200	———	11000 8000	0.31 1.40

TYPE	DESIGN	SOCKET CONN	CATHODE TYPE AND RATING			USED AS	PLATE SUPPLY VOLTS	GRID BIAS② VOLTS	SCREEN SUPPLY VOLTS	SCREEN CURRENT MILLIAMP.	PLATE CURRENT MILLIAMP.	A.C. PLATE RESISTANCE OHMS	TRANSCON-DUCTANCE (GRID-PLATE) µMHOS	AMPLI-FICATION FACTOR	LOAD FOR STATED POWER OUTPUT OHMS	POWER OUTPUT WATTS
			C.T.	VOLTS	AMP											
6A5-G	POWER AMPLIFIER TRIODE	6T	H	6.3	1.25	POWER OUTPUT AMPLIFIER	325	−68.0 FIXED BIAS −68.0 FIXED BIAS 850 OHMS SELF-BIAS RESISTOR			60.0 40.0 40.0	800 — —	5250 — —	4.2 — —	2500 PP 3000 PP 5000	1 TUBE 3.75 15 10 ⑫
6A6	TWIN-TRIODE AMPLIFIER	7B	H	6.3	0.8	CLASS B AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6N7-G.									
6A7	PENTAGRID CONVERTER ⑧	7C	H	6.3	0.3	CONVERTER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6A8									
6A8(G)(GT)	PENTAGRID CONVERTER ⑧	8A	H	6.3	0.3	CONVERTER	100 250	−1.5 −3.0	50 100	1.3 2.7	1.1 3.5	600000 360000	ANODE-GRID (#2): 250③MAX. VOLTS, 4.0 MA. OSCILLATOR-GRID (#1) RESISTOR ⑲ CONVERSION TRANSCOND., 550 µMHOS			
6AB5/6N5	ELECTRON-RAY TUBE	6R	H	6.3	0.15	VISUAL INDICATOR	PLATE & TARGET SUPPLY = 135 VOLTS. TRIODE PLATE RESISTOR = 0.25 MEG. TARGET CURRENT = 2.0 MA GRID BIAS, −10.0 VOLTS; SHADOW ANGLE, 0°; BIAS, 0 VOLTS; ANGLE, 90°; PLATE CURRENT, 0.5 MA.									
6AB6 (G)	DIRECT-COUPLED POWER AMPLIFIER	7AU	H	6.3	0.5	AMPLIFIER	250 250	0	—	—	4.0 34.0	VALUES FOR INPUT TRIODE VALUES FOR OUTPUT TRIODE			8000	3.5
6AB7/1853	TELEVISION AMPLIFIER PENTODE	8N	H	6.3	0.45	CLASS A AMPLIFIER	300	−3.0	200	3.2	12.5	700000	5000	—	—	—
6AC5-GT/G	HIGH-MU POWER AMPLIFIER TRIODE	6Q	H	6.3	0.4	CLASS B AMPLIFIER	250	0	—	—	5.0 ⑬	—	—	—	10000	8.0 ⑫
						DYNAMIC COUPLED AMPLIFIER WITH TYPE 6P5-G DRIVER	250	BIAS FOR BOTH 6AC5-G AND 6P5-G IS DEVELOPED IN COUPLING CIRCUIT, AVERAGE PLATE CURRENT OF DRIVER = 5.5 MILLIAMPERES AVERAGE PLATE CURRENT OF 6AC5-G = 32 MILLIAMPERES							7000	3.7
6AC6-G(GT)	TRIPLE-TWIN POWER AMPLIFIER	7W	H	6.3	1.1	POWER AMPLIFIER	180	0	—	—	OUTPUT 45 INPUT 7	18000	3000	54	3500	3.6
6AC7/1852	TELEVISION AMPLIFIER PENTODE	8N	H	6.3	0.45	CLASS A AMPLIFIER	300	CATHODE BIAS	150	2.5	10.0	750000	9000	CATHODE-BIAS RESISTOR, 160 OHMS		
6AD5-G	HIGH-MU TRIODE	6Q	H	6.3	0.3	HIGH-MU TRIODE AMPLIFIER	250	−2.0	—	—	0.9	66000	—	100	—	—
6AD6-G	ELECTRON-RAY TUBE	7AG	H	6.3	0.15	VISUAL INDICATOR	TARGET VOLTAGE 150−100	RAY CONTROL VOLTAGE 0 TO −50	—	—	—	—	—	—	—	—
6AD7-G	TRIODE A.F. POWER PENTODE	8AY	H	6.3	0.85	TRIODE UNIT AS CLASS A AMPLIFIER	250	−25.0	—	—	4.0	19000	325	6.0	—	—
						PENTODE UNIT AS CLASS A AMPLIFIER	PENTODE UNIT IS IDENTICAL WITH TYPE 6F6-G. 6F6-G AND 6AD7-G MAY BE USED TOGETHER AS COMBINED PHASE INVERTER AND PUSH-PULL POWER AMPLIFIER									
6AE5-GT/G	AMPLIFIER TRIODE	6Q	H	6.3	0.3	CLASS A AMPLIFIER	95	−15.0	—	—	7.0	3500	1200	4.2	—	—
6AE6-G	TWIN-PLATE CONTROL-TUBE	7AH	H	6.3	0.15	SHARP-CUTOFF TRIODE	250 250	−1.5 −9.5	—	—	4.5 0.01	35000	950	33.0	CONTROL TUBE FOR TWIN-ELECTRON-RAY INDICATOR TUBES, SUCH AS 6AF6-G	
						REMOTE CUTOFF TRIODE	250 250	−1.5 −35.0	—	—	6.5 0.01	25000	1000	25.0		
6AE7-GT	DOUBLE-DRIVER TRIODE	7AX	H	6.3	0.5	DRIVER	250	−13.5 ㊳	—	—	5.0	9300	1500	14	—	—
6AF5-G (GT)	AMPLIFIER TRIODE	6Q	H	6.3	0.3	CLASS A AMPLIFIER	180	−18.0	—	—	7.0	4900	1500	7.4	—	—

6AF6-G	ELEC.-RAY TUBE TWIN-INDI.-TYPE	7AG	H	6.3	0.15	VISUAL INDICATOR	TARGET VOLTAGE, 100 VOLTS. CONTROL ELECTRODE VOLTAGE, 0 VOLTS; SHADOW ANGLE, 100°; TARGET CURRENT 0.9 MA., CONTROL-ELECTRODE VOLTAGE, 60 VOLTS; ANGLE, 0°. TARGET VOLTAGE, 135 VOLTS. CONTROL ELECTRODE VOLTAGE, 0 VOLTS; SHADOW ANGLE, 100°; TARGET CURRENT, 1.5 MA. CONTROL-ELECTRODE VOLTAGE, 81 VOLTS; ANGLE, 0°								
6AF7-G	ELEC.-RAY TUBE TWIN-INDI. TYPE	8AG	H	6.3	0.3	VISUAL INDICATOR									
6AG6 (G)	POWER AMPLIFIER PENTODE	7S	H	6.3	1.25	CLASS A AMPLIFIER	250	−6.0	250	6.0	32	—	10000	—	8500	3.75
6AG7	VIDEO POWER AMPLIFIER PENTODE	8Y	H	6.3	0.65	CLASS A AMPLIFIER	300	−2.0	125	7.0	28.0	100000	7700	—	3500	PEAK-TO-PEAK VOLTS OUTPUT 140 APPROX.
6AH7-GT	TWIN-TRIODE AMPLIFIER	8BE	H	6.3	0.3	AMPLIFIER	250	−9.0	—	—	12	6600	2400	16	—	—
6AL6-G	BEAM POWER AMPLIFIER	6AM	H	6.3	0.9	CLASS A AMPLIFIER	250	−14.0	250	5.0	72.0	22500	6000	—	2500	6.5
6B4-G	POWER AMPLIFIER TRIODE	5S	F	6.3	1.0	CLASS A AMPLIFIER	325 325	CATHODE BIAS 850 OHMS (13) −68 VOLTS, FIXED BIAS (13)			80.0 (13) 80.0 (13)	—	—	—	5000 5000	10.0 (12) 15.0 (12)
6B5	DIRECT-COUPLED POWER AMPLIFIER	6AS	H	6.3	0.8	CLASS A AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6N6-G.									
6B6-G	DUPLEX-DIODE HIGH-MU TRIODE	7V	H	6.3	0.3	TRIODE UNIT AS AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6SQ7.									
6B7	DUPLEX-DIODE PENTODE	7D	H	6.3	0.3	PENTODE UNIT AS AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6B8-G.									
6B8 (G)	DUPLEX-DIODE PENTODE	8E	H	6.3	0.3	PENTODE UNIT AS R.F. AMPLIFIER	100 250	−3.0 −3.0	100 125	1.7 2.3	5.8 9.0	300000 600000	950 1125	—	—	—
						PENTODE UNIT AS A.F. AMPLIFIER	90 (18) 300 (18)	CATHODE BIAS, 3500 OHMS. SCREEN RESISTOR = 1.1 MEG. CATHODE BIAS, 1600 OHMS. SCREEN RESISTOR = 1.2 MEG.				GRID RESISTOR, (10) 0.5 MEGOHM		GAIN PER STAGE = 55 GAIN PER STAGE = 79		
6C5 (G) (GT)	DETECTOR AMPLIFIER TRIODE (1)	6Q	H	6.3	0.3	CLASS A AMPLIFIER	250	−8.0	—	—	8.0	10000	2000	20	—	—
						BIAS DETECTOR	250	−17.0 APPROXIMATE PLATE CURRENT TO BE ADJUSTED TO 0.2 MILLIAMPERE WITH NO SIGNAL.								
6C6	TRIPLE-GRID DETECTOR AMPLIFIER	6F	H	6.3	0.3	AMPLIFIER DETECTOR	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6J7.									
6C7	DUPLEX-DIODE TRIODE	7G	H	6.3	0.3	TRIODE UNIT AS AMPLIFIER	250	−9.0	—	—	5.5	—	1250	20	—	—
6C8-G	TWIN-TRIODE AMPLIFIER	8G	H	6.3	0.3	EACH UNIT AS AMPLIFIER	250	−4.5	—	—	3.2	22500	1600	36	—	—
6D6	TRIPLE-GRID SUPER CONTROL AMPLIFIER	6F	H	6.3	0.3	AMPLIFIER MIXER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6U7-G.									
6D7	TRIPLE-GRID DETECTOR AMPLIFIER	7H	H	6.3	0.3	AMPLIFIER DETECTOR	FOR OTHER CHARACTERISTICS REFER TO TYPE 6J7									
6D8-G	PENTAGRID CONVERTER (8)	8A	H	6.3	0.15	CONVERTER	135 250	−3.0 −3.0	67.5 100	—	—	600000 400000	ANODE-GRID (#2): 250 (3) MAX. VOLTS, 4.3 MA, OSCILLATOR-GRID (#1) RESISTOR (19) CONVERSION TRANSCONDUCTANCE, 550 µMHOS			
6E5	ELECTRON-RAY TUBE	6R	H	6.3	0.3	VISUAL INDICATOR	PLATE & TARGET SUPPLY = 100 VOLTS, TRIODE PLATE RESISTOR = 0.5 MEG. TARGET CURRENT = 1.0 MA. GRID BIAS, −3.3 VOLTS; SHADOW ANGLE, 0°. BIAS, 0 VOLTS; ANGLE, 90°; PLATE CURRENT, 0.19 MA. PLATE & TARGET SUPPLY = 250 VOLTS. TRIODE PLATE RESISTOR = 1.0 MEG. TARGET CURRENT = 4.0 MA. GRID BIAS, −8.0 VOLTS; SHADOW ANGLE, 0°. BIAS, 0 VOLTS; ANGLE, 90°; PLATE CURRENT, 0.24 MA.								
6E6	TWIN TRIODE	7B	H	6.3	0.6	PUSH-PULL CLASS A AMPLIFIER	180 250	−20.0 −27.5	—	—	11.5 18.0	4300 3500	1400 1700	6 6	15000 14000	0.75 1.6
6E7	TRIPLE-GRID SUPER-CONTROL AMPLIFIER	7H	H	6.3	0.3	AMPLIFIER MIXER	FOR OTHER CHARACTERISTICS REFER TO TYPE 6U7-G									
6E8 (G)	TRIODE-HEXODE CONVERTER	8O	H	6.3	0.3	OSCILLATOR-MIXER	150 250	0 −2.0	VALUES FOR TRIODE UNIT VALUES FOR HEXODE UNIT			— 1250000	2800 —	CONVERSION TRANSCONDUCTANCE 2800 MICROMHOS		

TYPE	DESIGN	SOCKET CONN.	CATHODE TYPE AND RATING C.T.	VOLTS	AMP.	USED AS	PLATE SUPPLY VOLTS	GRID BIAS (2) VOLTS	SCREEN SUPPLY VOLTS	SCREEN CURRENT MILLIAMP.	PLATE CURRENT MILLIAMP.	A.C. PLATE RESISTANCE OHMS	TRANSCONDUCTANCE (GRID-PLATE) μMHOS	AMPLIFICATION FACTOR	LOAD FOR STATED POWER OUTPUT OHMS	POWER OUTPUT WATTS
6F5 (G) (GT)	HIGH-MU TRIODE	5M	H	6.3	0.3	AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6SF5									
6F6 (G) (GT)	POWER AMPLIFIER PENTODE	7S	H	6.3	0.7	PENTODE CLASS A AMPLIFIER	250 285	−16.5 −20.0	250 285	6.5 7.0	34.0 38.0	80 000 78000	2500 2550	—	7000 7000	3.2 4.8
						TRIODE (21) CLASS A AMPLIFIER	250	−20.0	—	—	31.0	2600	2600	6.8	4000	0.85
						PENTODE PUSH-PULL CLASS A AMPLIFIER	315 315	CATH. BIAS −24.0	285 285	12.0 (13) 12.0 (13)	62.0 (13) 62.0 (13)	CATH. BIAS RESISTOR, 320 OHMS (13) —	— —	— —	10 000 10000	10.5 (12) 11.0 (12)
						PENTODE PUSH-PULL CLASS AB_2 AMPLIFIER	375 375	CATH. BIAS −26.0	250 250	8.0 (13) 5.0 (13)	54.0 (13) 34.0 (13)	CATHODE BIAS RESISTOR, 340 OHMS (13) —	— —	— —	10000 10000	19.0 (12) 18.5 (12)
						TRIODE PUSH-PULL (21) CLASS AB_2 AMPLIFIER	350 350	CATH. BIAS −38.0	—	—	50.0 (13) 48.0 (13)	CATHODE BIAS RESISTOR, 730 OHMS (13) —	— —	— —	10000 6000	9.0 (12) 13.0 (12)
6F7	TRIODE PENTODE	7E	H	6.3	0.3	TRIODE UNIT AS CLASS A AMPLIFIER	100	{−3.0 MIN.}	—	—	3.5	16000	500	8.0	—	—
						PENTODE UNIT AS CLASS A AMPLIFIER	100 250	{−3.0 MIN.}	100 100	1.6 1.5	6.3 6.5	290 000 850 000	1050 1100	—	—	—
						PENTODE UNIT AS MIXER	250	−10.0	100	0.6	2.8	OSCILLATOR PEAK VOLTS = 7.0 CONVERSION TRANSCONDUCTANCE = 300 MICROMHOS				
6F8-G	TWIN TRIODE AMPLIFIER	8G	H	6.3	0.6	EACH UNIT AS AMPLIFIER	90 250	0 −8.0	—	—	10.0 9.0	6700 7700	3000 2600	20 20	—	—
6G5/6U5	ELECTRON-RAY TUBE	REFER TO 6U5 DATA														
6G6-G	POWER AMPLIFIER PENTODE	7S	H	6.3	0.15	PENTODE CLASS A AMPLIFIER	135 180	−6.0 −9.0	135 180	2.0 2.5	11.5 15.0	170000 175 000	2100 2300	—	12000 10000	0.6 1.1
						TRIODE (21) CLASS A AMPLIFIER	180	−12.0	—	—	11.0	4750	2000	9.5	12000	0.25
6H4-GT	DIODE	5AF	H	6.3	0.15	DETECTOR RECTIFIER	100	—	—	—	4.0	—	—	—	—	—
6H5	ELECTRON-RAY TUBE	6R	H	6.3	0.3	VISUAL INDICATOR	PLATE & TARGET SUPPLY = 250 VOLTS. TARGET CURRENT = 4.5 MA. GRID BIAS, −22.0 VOLTS; SHADOW ANGLE, 0°. BIAS, 0 VOLTS; SHADOW ANGLE, 90°									
6H6-GT/6H6-G	TWIN DIODE	7Q	H	6.3	0.3	DETECTOR RECTIFIER	MAXIMUM A.C. VOLTAGE PER PLATE . . . . 150 VOLTS, RMS MAXIMUM D.C. OUTPUT CURRENT . . . . 8 MILLIAMPERES									
6H8-G	DUPLEX-DIODE PENTODE	8E	H	6.3	0.3	PENTODE UNIT AS AMPLIFIER	250	−2.0	—	—	8.5	650 000	2400	—	—	—
6J5 (G) (GT)	DETECTOR-AMPLIFIER-TRIODE	6Q	H	6.3	0.3	CLASS A AMPLIFIER	90 250	0 −8.0	—	—	10.0 9.0	6700 7700	3000 2600	20 20	—	—
6J7 (G) (GT)	TRIPLE-GRID DETECTOR AMPLIFIER	7R	H	6.3	0.3	PENTODE CLASS A R.F. AMPLIFIER	100 250	−3.0 −3.0	100 100	0.5 0.5	2.0 2.0	1000000 1000000 +	1185 1225	—	—	—
						PENTODE CLASS A A.F. AMPLIFIER	90 (18) 300 (18)	CATHODE BIAS, 2600 OHMS. SCREEN RESISTOR = 1.2 MEG. } GRID RESISTOR (10) 0.5 MEGOHM } GAIN PER STAGE = 85 " " 1200 " " " GAIN PER STAGE = 140								
						PENTODE BIAS DETECTOR	250	−4.3	100	CATHODE CURRENT 0.43 MA.		—	PLATE RESISTOR, 500 000 OHMS GRID RESISTOR, (10) 250 000 OHMS			
						TRIODE (15) CLASS A AMPLIFIER	180 250	−5.3 −8.0	—	—	5.3 6.5	11000 10500	1800 1900	20 20	—	—

							Ep	Eg	Esg	Isg	Ip	Rp	Gm	µ		
6J8-G	TRIODE-HEPTODE CONVERTER	8H	H	6.3	0.3	TRIODE SECTION AS OSCILLATOR	100 250	50000 Ω 50000 Ω	———	———	3.0 5.0	8750	1600	14	———	———
												PLATE VOLTAGE APPLIED THROUGH 20000 OHM RESISTOR				
						HEPTODE SECTION AS MIXER	100 250	−3.0 −3.0	100 100	3.0 2.9	1.4 1.3	900000 4000000	250 290			
6K5-G (GT)	HIGH-MU TRIODE	5U	H	6.3	0.3	CLASS A AMPLIFIER	100 250	−1.5 −3.0	———	———	0.35 1.1	78000 50000	900 1400	70 70	———	———
6K6-GT/G	POWER AMPLIFIER PENTODE	7S	H	6.3	0.4	CLASS A AMPLIFIER	100 180 250 315	−7.0 −13.5 −18.0 −21.0	100 180 250 250	1.6 3.0 5.5 4.0	9.0 18.5 32.0 25.5	104000 81000 68000 75000	1500 1850 2300 2100	———	12000 9000 7600 9000	0.35 1.50 3.40 4.50
6K7 (G)(GT)	TRIPLE-GRID SUPER-CONTROL AMPLIFIER	7R	H	6.3	0.3	CLASS A AMPLIFIER	90 250	−3.0 −3.0	90 125	1.3 2.6	5.4 10.5	300000 600000	1275 1650	———	———	———
						MIXER IN SUPERHETERODYNE	250	−10.0	100	OSCILLATOR PEAK VOLTS = 7.0						
6K8(G) (GT)	TRIODE-HEXODE CONVERTER	8K	H	6.3	0.3	TRIODE UNIT AS OSCILLATOR	100	TRIODE-GRID RESISTOR 50000 Ω			3.8	TRIODE-GRID & HEXODE-GRID CURRENT = 0.15 MA.				
						HEXODE UNIT AS MIXER	100 250	−3.0 −3.0	100 100	6.2 6.0	2.3 2.5	400000 600000	CONVERSION TRANSCOND., 325 µMHOS. CONVERSION TRANSCOND., 350 µMHOS.			
6L5-G	DETECTOR AMPLIFIER TRIODE	6Q	H	6.3	0.15	CLASS A AMPLIFIER	135 250	−5.0 −9.0	———	———	3.5 8.0	11200 9000	1500 1900	17 17	———	———
6L6 (G)	BEAM POWER AMPLIFIER	7AC	H	6.3	0.9	SINGLE-TUBE CLASS A AMPLIFIER	250 250	−14.0 CATH. BIAS	250 250	5.0 5.4	72.0 75.0	22500	6000	———	2500 2500	6.5 6.5
												CATHODE BIAS RESISTOR, 170 OHMS.				
						PUSH-PULL CLASS A AMPLIFIER	270 270	−17.5 CATH. BIAS	270 270	11.0 (13) 11.0 (13)	134.0 (13) 134.0 (13)	23500	5700	———	5000 5000	17.5 (12) 18.5 (12)
												CATHODE BIAS RESISTOR 125 OHMS. (13)				
						PUSH-PULL CLASS AB_1 AMPLIFIER	360 360	−22.5 CATH. BIAS	270 270	5.0 (13) 5.0 (13)	88.0 (13) 88.0 (13)	———	———	———	6600 9000	26.5 (12) 24.5 (12)
												CATHODE BIAS RESISTOR, 250 OHMS (13)				
						PUSH-PULL CLASS AB_2 AMPLIFIER	360 360	−18.0 −22.5	225 270	3.5 (13) 5.0 (13)	78.0 (13) 88.0 (13)	———	———	———	6000 3800	31.0 (12) 47.0 (12)
						SINGLE TRIODE (21) CLASS A AMPLIFIER	250 250	−20.0 CATH. BIAS	———	———	40.0 40.0	1700	4700	8.0	5000 6000	1.4 1.3
												CATHODE BIAS RESISTOR, 490 OHMS				
6L7 (G)	PENTAGRID MIXER (9) AMPLIFIER	7T	H	6.3	0.3	MIXER IN SUPERHETERODYNE	250	−3.0	100	7.1	2.4	OSCILLATOR-GRID (#3) BIAS, −10 VOLTS. GRID #3 PEAK SWING, 12 VOLTS MINIMUM. CONVERSION TRANSCOND., 375 MICROMHOS				
						CLASS A AMPLIFIER	250	−3.0 (24)	100	6.5	5.3	600000	1100	———	———	———
6M6-G	POWER AMPLIFIER PENTODE	7S	H	6.3	1.2	CLASS A AMPLIFIER	250	−6.0	250	4.0	36.0	———	9500	———	7000	4.40
6M7-G	R.F. AMPLIFIER PENTODE	7R	H	6.3	0.3	AMPLIFIER	250	−2.5	125	2.8	10.5	900000	3400	———	———	———
6M8-G (GT)	DIODE-TRIODE PENTODE	8AU	H	6.3	0.6	TRIODE UNIT AS A.F. AMPLIFIER	100	−1.0	———	———	0.5	91000	1100	———	———	———
						PENTODE UNIT AS R.F. AMPLIFIER	100	−3.0	———	———	8.5	200000	1900	———	———	———
6N5	ELECTRON-RAY TUBE	6R	H	6.3	1.15	VISUAL INDICATOR	SUPERSEDED BY TYPE 6AB5/6N5									
6N6-G	DIRECT-COUPLED POWER AMPLIFIER	7AU	H	6.3	0.8	CLASS A AMPLIFIER	OUTPUT TRIODE: PLATE VOLTS, 300, PLATE MA., 42, LOAD, 7000 OHMS INPUT TRIODE: PLATE VOLTS, 300; GRID VOLTS, 0; A.F. SIGNAL VOLTS (RMS), 15, PLATE MA., 9.0									4.0
6N7-GT/G	TWIN-TRIODE AMPLIFIER	8B	H	6.3	0.8	CLASS A AMPLIFIER (AS DRIVER) (11)	250 294	−5.0 −6.0	———	———	6.0 7.0	11300 11000	3100 3200	35 35	20000 OR MORE	EXCEEDS 0.4
						CLASS B AMPLIFIER	300 300	0 0	———	———	35.0 35.0	POWER OUTPUT IS FOR ONE TUBE AT STATED PLATE-TO-PLATE LOAD.			8000 8000	10.0 10.0

TYPE	DESIGN	SOCKET CONN.	CATHODE TYPE AND RATING C.T.	VOLTS	AMP.	USED AS	PLATE SUPPLY VOLTS	GRID BIAS (2) VOLTS	SCREEN SUPPLY VOLTS	SCREEN CURRENT MILLIAMP.	PLATE CURRENT MILLIAMP.	A.C. PLATE RESISTANCE OHMS	TRANSCONDUCTANCE (GRID-PLATE) µMHOS	AMPLIFICATION FACTOR	LOAD FOR STATED POWER OUTPUT OHMS	POWER OUTPUT WATTS
6P5-GT/G	DETECTOR AMPLIFIER TRIODE	6Q	H	6.3	0.3	CLASS A AMPLIFIER	100 250	−5.0 −13.5	—	—	2.5 5.0	12000 9500	1150 1450	13.8 13.8	—	—
							90 (17) 300 (17)	CATHODE BIAS, 6500 OHMS. CATHODE BIAS, 6400 OHMS.	GRID RESISTOR,(10) 0.25 MEGOHM				GAIN PER STAGE = 9 GAIN PER STAGE = 10			
						BIAS DETECTOR	250	−20.0 APPROX.	—	—	PLATE CURRENT ADJUSTED TO 0.2 MILLIAMPERE WITH NO SIGNAL.					
6P7-G	TRIODE PENTODE	7U	H	6.3	0.3	TRIODE UNIT AS OSCILLATOR	100	−3.0	—	—	2.4	16 200	525	8.5	D.C. GRID CURRENT = 0.15 MILLIAMPERE	
						PENTODE UNIT AS MIXER	250	−10.0	100	0.6	2.8	2000000	CONVERSION TRANSCOND., 300 µMHOS OSCILLATOR PEAK VOLTS = 7.0			
6P8-G	TRIODE HEXODE	8K	H	6.3	0.8	TRIODE UNIT AS OSCILLATOR	100	TRIODE-GRID RESISTOR = 50000 Ω			2.2	—	—	—	—	—
						PENTODE UNIT AS MIXER	250	−2.4	75 (26)	1.4	1.5	—	650	—	—	—
6Q6-G	DIODE-TRIODE	6Y	H	6.3	0.15	TRIODE UNIT AS CLASS A AMPLIFIER	250	−3.0	—	—	1.2	—	1050	65	—	—
6Q7(G)(GT)	DUPLEX-DIODE HIGH-MU TRIODE	7V	H	6.3	0.3	TRIODE UNIT AS CLASS A AMPLIFIER	100 250	−1.5 −3.0	—	—	0.35 1.1	87500 58000	800 1200	70 70	—	—
							90 (18) 300 (18)	CATHODE BIAS, 7600 OHMS. CATHODE BIAS, 3000 OHMS.	GRID RESISTOR,(10) 0.5 MEGOHM				GAIN PER STAGE = 32 GAIN PER STAGE = 45			
6R6-G	REMOTE CUTOFF R.F. PENTODE	6AW	H	6.3	0.3	CLASS A AMPLIFIER	250	−3.0	100	1.7	7.0	800000	1450	—	—	—
6R7(G) (GT)	DUPLEX-DIODE TRIODE	7V	H	6.3	0.3	TRIODE UNIT AS CLASS A AMPLIFIER	250	−9.0	—	—	9.5	8500	1900	16	—	—
							90 (17) 300 (17)	CATHODE BIAS, 4400 OHMS. CATHODE BIAS, 3800 OHMS.	GRID RESISTOR,(10) 0.25 MEGOHM				GAIN PER STAGE = 10 GAIN PER STAGE = 10			
6S5	ELECTRON-RAY TUBE	6R	H	6.3	0.3	VISUAL INDICATOR	FOR OTHER CHARACTERISTICS REFER TO TYPE 6X6 (G)									
6S7 (G)	TRIPLE-GRID AMPLIFIER	7R	H	6.3	0.15	CLASS A AMPLIFIER	135 250	−3.0 −3.0	67.5 100	0.9 2.0	3.7 8.5	1000 000 1000 000	1250 1750	—	—	—
6SA7 (GT)	PENTAGRID CONVERTER (26)	8R	H	6.3	0.3	CONVERTER	100 250	—	100 100	8.5 8.5	3.3 3.5	500000 1000000	GRID #1 RESISTOR, 20000 OHMS. CONVERSION TRANSCOND., 450 µMHOS			
6SC7	TWIN-TRIODE AMPLIFIER	8S	H	6.3	0.3	EACH UNIT AS AMPLIFIER	250	−2.0	—	—	2.0	53000	1325	70	—	—
6SD7-GT	R.F. AMPLIFIER PENTODE	8N	H	6.3	0.3	CLASS A AMPLIFIER	250	−2.0	100	1.9	6.0	1000 000	3600	—	—	—
6SE7-GT	R.F. AMPLIFIER PENTODE	8N	H	6.3	0.3	CLASS A AMPLIFIER	250	−1.5	100	1.5	4.5	1100000	3400	—	—	—
6SF5 (GT)	HIGH-MU TRIODE	6AB	H	6.3	0.3	CLASS A AMPLIFIER	100 250	−1.0 −2.0	—	—	0.4 0.9	85000 66000	1150 1500	100 100	—	—
							90 (18) 300 (18)	CATHODE BIAS, 8800 OHMS. CATHODE BIAS, 3200 OHMS.	GRID RESISTOR,(10) 0.5 MEGOHM				GAIN PER STAGE = 43 GAIN PER STAGE = 63			
6SF7	DIODE-SUPER-CONTROL AMPLIFIER PENTODE	7AZ	H	6.3	0.3	PENTODE UNIT AS CLASS A AMPLIFIER	100 250	−1.0 −1.0	100 100	3.4 3.3	12.0 12.4	200000 700000	1975 2050	—	—	—

6SG7	H.F. AMPLIFIER PENTODE	8BC	H	6.3	0.3	CLASS A AMPLIFIER	100 250	−11.5 −17.5	100 150	3.2 3.4	8.2 9.2	250000 1000000+	4100 4000	—	—	—
6SJ7 (GT)	TRIPLE-GRID DETECTOR AMPLIFIER	8N	H	6.3	0.3	CLASS A AMPLIFIER	100 250	−3.0 −3.0	100 100	0.9 0.8	2.9 3.0	700000 1500000	1575 1650	—	—	—
							90 (18) 300 (18)	CATHODE BIAS, 1700 OHMS. CATHODE BIAS, 860 OHMS.	GRID RESISTOR, (10) 0.5 MEGOHM				GAIN PER STAGE = 93 GAIN PER STAGE = 167			
6SK7 (GT)	TRIPLE-GRID AMPLIFIER.	8N	H	6.3	0.3	CLASS A AMPLIFIER	100 250	−1.0 −3.0	100 100	4.0 2.6	13.0 9.2	120000 800000	2350 2000	—	—	—
6SL7-GT	TWIN-TRIODE AMPLIFIER	8BD	H	6.3	0.3	CLASS A AMPLIFIER (39)	250	−2.0	—	—	2.3	44000	1600	70	—	—
6SN7-GT	TWIN-TRIODE AMPLIFIER	8BD	H	6.3	0.3	CLASS A AMPLIFIER (39)	90 250	0 −8.0 (40)	—	—	10.0 9.0	6700 7700	3000 2600	20 20	—	—
6SQ7-GT/G	DUPLEX-DIODE HIGH-MU TRIODE	8Q	H	6.3	0.3	TRIODE UNIT AS CLASS A AMPLIFIER	250	−2.0	—	—	0.9	91000	1100	100	—	—
							90 (18) 300 (18)	CATHODE BIAS, 11000 OHMS CATHODE BIAS, 3900 OHMS	GRID RESISTOR, (10) 0.5 MEGOHM				GAIN PER STAGE = 40 GAIN PER STAGE = 53			
6SR7	DUPLEX-DIODE TRIODE	8Q	H	6.3	0.3	TRIODE UNIT AS CLASS A AMPLIFIER	250	−9.0	—	—	9.5	8500	1900	16	10000	0.3
6SS7	TRIPLE-GRID SUPER-CONTROL AMPLIFIER	8N	H	6.3	0.15	CLASS A AMPLIFIER	100 250	−1.0 −3.0	100 100	3.1 2.0	12.2 9.0	120000 1000000	1930 1850	—	—	—
6T5	SUPERSEDED BY TYPE 6U5/6G5															
6T6	R.F. AMPLIFIER PENTODE	6Z	H	6.3	0.45	CLASS A AMPLIFIER	250	−1.0	100	2.0	10.0	1000000	5500	—	—	—
6T7-G	DUPLEX-DIODE HIGH-MU TRIODE	7V	H	6.3	0.15	TRIODE UNIT AS CLASS A AMPLIFIER	250	−3.0	—	—	1.2	62000	1050	65	—	—
							90 (18) 300 (18)	CATHODE BIAS, 8300 OHMS CATHODE BIAS, 4580 OHMS	GRID RESISTOR, (10) 0.5 MEGOHM				GAIN PER STAGE = 30 GAIN PER STAGE = 40			
6U5/6G5	ELECTRON-RAY TUBE	6R	H	6.3	0.3	VISUAL INDICATOR	PLATE & TARGET SUPPLY = 100 VOLTS. TRIODE PLATE RESISTOR = 0.5 MEG. TARGET CURRENT = 1.0 MA. GRID BIAS, −8 VOLTS; SHADOW ANGLE, 0°. BIAS, 0 VOLTS; ANGLE, 90°; PLATE CURRENT, 0.19 MA.									
							PLATE & TARGET SUPPLY = 250 VOLTS. TRIODE PLATE RESISTOR = 1.0 MEG. TARGET CURRENT = 4.0 MA. GRID BIAS, −22 VOLTS; SHADOW ANGLE, 0°. BIAS, 0 VOLTS; ANGLE, 90°; PLATE CURRENT, 0.24 MA.									
6U6-GT	BEAM POWER AMPLIFIER	7AC	H	6.3	0.75	CLASS A AMPLIFIER	200	−14.0	135	3.0	56.0	20000	6200	—	3000	5.5
6U7-G	TRIPLE-GRID SUPER-CONTROL AMPLIFIER	7R	H	6.3	0.3	CLASS A AMPLIFIER	100 250	−3.0 −3.0	100 100	2.2 2.0	8.0 8.2	250000 800000	1500 1600	—	—	—
						MIXER IN SUPERHETERODYNE	100 250	−10.0 −10.0	100 100	—	—	OSCILLATOR PEAK VOLTS = 7.0				
6V6(GT/G)	BEAM POWER AMPLIFIER	7AC	H	6.3	0.45	SINGLE-TUBE CLASS A AMPLIFIER	180 315	−8.5 −13.0	180 225	3.0 2.2	29.0 34.0	58000 77000	3700 3750	—	5500 8500	2.0 5.5
						PUSH-PULL CLASS AB_1 AMPLIFIER	250 300	−15.0 −20.0	250 300	5.0 (13) 5.0 (13)	70.0 (13) 78.0 (13)	—	—	—	10000 8000	8.5 (12) 13.0 (12)
6V7-G	DUPLEX-DIODE TRIODE	7V	H	6.3	0.3	TRIODE UNIT AS CLASS A AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 85									
6W5(G)	FULL-WAVE RECTIFIER	6S	H	6.3	0.9	WITH CONDENSER-INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS) 325				MAX. D.C. OUTPUT MA., 90					
						WITH CHOKE INPUT FILTER	MAX. VOLTS PER PLATE (RMS) 450				MAX. D.C. OUTPUT MA., 90			VOLTAGE DROP AT 90 MA., 24 VOLTS		
6W6-GT	BEAM POWER AMPLIFIER	7AC	H	6.3	1.25	CLASS A AMPLIFIER	135	−9.5	135	2.5	58.0	—	9000	150	2000	3.5
6W7-G	TRIPLE-GRID DETECTOR AMPLIFIER	7R	H	6.3	0.15	CLASS A AMPLIFIER	250	−3.0	100	0.5	2.0	1500000	1225	—	—	—

TYPE	DESIGN	SOCKET CONN.	CATHODE TYPE AND RATING C.T.	VOLTS	AMP.	USED AS	PLATE SUPPLY VOLTS	GRID BIAS ② VOLTS	SCREEN SUPPLY VOLTS	SCREEN CURRENT MILLIAMP.	PLATE CURRENT MILLIAMP.	A.C. PLATE RESISTANCE OHMS	TRANSCON-DUCTANCE (GRID-PLATE) µMHOS	AMPLI-FICATION FACTOR	LOAD FOR STATED POWER OUTPUT OHMS	POWER OUTPUT WATTS
6X5(GT/G)	FULL-WAVE RECTIFIER	6S	H	6.3	0.6	WITH CONDENSER INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 325 MAX. PEAK INVERSE VOLTS, 1250				MAX. D.C. OUTPUT MA., 70 MAX. PEAK PLATE MA., 210			MIN. TOTAL EFFECTIVE SUPPLY IMPEDANCE PER PLATE, 150 OHMS		
						WITH CHOKE INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 450 MAX. PEAK INVERSE VOLTS, 1250				MAX. D.C. OUTPUT MA., 70 MAX. PEAK PLATE MA., 210			MIN. VALUE OF INPUT CHOKE, 8 HENRIES		
6X6 (G)	ELECTRON-RAY TUBE	7AL	H	6.3	0.3	VISUAL INDICATOR	TARGET 250	−8.0 0	VANE GRID 135	TARGET CURRENT, 0 MA. TARGET CURRENT, 2 MA.		VALUES FOR 0° ILLUMINATION ANGLE VALUES FOR 300° ILLUMINATION ANGLE				
6Y5	FULL-WAVE RECTIFIER ④	6J	H	6.3	0.8	RECTIFIER	MAX. A.C. VOLTS PER PLATE (RMS), 350 MAX. PEAK INVERSE VOLTS, 1500				MAX. D.C. OUTPUT MA., 50 MAX. PEAK PLATE MA., 200					
6Y6-G (GT)	BEAM POWER AMPLIFIER	7AC	H	6.3	1.25	SINGLE-TUBE CLASS A AMPLIFIER	135 200	−13.5 −14.0	135 135	3.5 2.2	58.0 61.0	9300 18300	7000 7100	—	2000 2600	3.6 6.0
6Y7-G	TWIN-TRIODE AMPLIFIER	8B	H	6.3	0.6	CLASS B AMPLIFIER	180 250	0 0	—	—	7.6 10.6	—	—	—	7000 14000	5.5 8.0
6Z4/84	REFER TO TYPE 84 DATA															
6Z5	FULL-WAVE RECTIFIER	6K	H	12.6 6.3	0.4 0.8	RECTIFIER	MAX. A.C. VOLTS PER PLATE (RMS), 230. MAX. D.C. OUTPUT MA., 60 MAX. PEAK INVERSE VOLTS, 1500.									
6Z7-G	TWIN-TRIODE AMPLIFIER	8B	H	6.3	0.3	CLASS B AMPLIFIER	135 180	0 0	—	—	POWER OUTPUT IS FOR ONE TUBE AT STATED PLATE-TO-PLATE LOAD.				9000 12000	2 5 4 2
6ZY5-G	FULL-WAVE RECTIFIER	6S	H	6.3	0.3	WITH CONDENSER-INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 325. MAX. PEAK INVERSE VOLTS, 1250				MAX. D.C. OUTPUT MA., 40 MAX. PEAK PLATE MA., 120			MIN. TOTAL EFFECTIVE SUPPLY IMPEDANCE PER PLATE, 225 Ω		
						WITH CHOKE-INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 450 MAX. PEAK INVERSE VOLTS, 1250				MAX. D.C. OUTPUT MA., 40 MAX. PEAK PLATE MA., 120			MINIMUM VALUE OF INPUT CHOKE, 13.5 HENRIES.		
7A4	DETECTOR AMPLIFIER TRIODE	5AC	H	6.3	0.3	CLASS A AMPLIFIER	90 250	0 −8.0	—	—	10.0 9.0	6700 7700	3000 2600	20 20	—	—
7A5	POWER AMPLIFIER PENTODE	6AA	H	6.3	0.75	CLASS A AMPLIFIER	110 125	−7.5 −9.0	110 125	3.0 3.3	40.0 44.0	14000 17000	5800 6000	—	2500 2700	1.5 2.2
7A6	TWIN DIODE	7AJ	H	6.3	0.15	DETECTOR RECTIFIER	MAXIMUM A.C VOLTAGE PER PLATE . . . . 150 VOLTS, RMS MAXIMUM D.C. OUTPUT CURRENT . . . . 10 MILLIAMPERES									
7A7-LM	TRIPLE-GRID SUPER-CONTROL AMPLIFIER	8V	H	6.3	0.3	CLASS A AMPLIFIER	250	{−3.0 MIN.}	100	2.0	8.6	800000	2000	—	—	—
7A8	OCTODE CONVERTER	8U	H	6.3	0.15	CONVERTER	250	{−3.0 MIN.}	100	3.2	3.0	700000	ANODE-GRID (#2): 200 ③ MAX. VOLTS, 4.2 MA. OSCILLATOR-GRID (#1) RESISTOR, 50000 Ω CONV. TRANSCOND., 550 µMHOS.			
7B4	HIGH-MU TRIODE	5AC	H	6.3	0.3	CLASS A AMPLIFIER	100 250	−1.0 −2.0	—	—	0.5 0.9	85000 66000	1175 1500	100 100	—	—
7B5-LT	POWER AMPLIFIER PENTODE	6AE	H	6.3	0.4	CLASS A AMPLIFIER	100 250 315	−7.0 −18.0 −21.0	100 250 250	1.6 5.5 4.0	9.0 32.0 25.5	104000 68000 75000	1500 2300 2100	—	12000 7600 9000	0.35 3.4 4.5
7B6-LM	DUPLEX-DIODE HIGH-MU TRIODE	8W	H	6.3	0.3	TRIODE UNIT AS AMPLIFIER	100 250	−1.0 −2.0	—	—	0.25 0.9	132000 91000	760 1100	100 100	—	—

7B7	TRIPLE-GRID SUPER-CONTROL AMPLIFIER	8V	H	6.3	0.15	CLASS A AMPLIFIER	250	{−3.0 MIN.}	100	2.0	8.5	700000	1700	—	—	—
7B8-LM	PENTAGRID CONVERTER	8X	H	6.3	0.3	CONVERTER	100 250	−1.5 −3.0	50 100	1.3 2.7	1.1 3.5	600000 360000	360 550	ANODE-GRID (#2) 250③MAX. VOLTS, 4.0 MA. OSCILLATOR—GRID (#1) RESISTOR, 50000 OHMS		
7C5-LT	BEAM POWER AMPLIFIER	6AA	H	6.3	0.45	CLASS A AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6V6									
7C6	DUPLEX-DIODE HIGH-MU TRIODE	8W	H	6.3	0.15	TRIODE UNIT AS CLASS A AMPLIFIER	250	−1.0	—	—	1.3	100000	1000	100	—	—
7C7	TRIPLE-GRID DETECTOR AMPLIFIER	8V	H	6.3	0.15	CLASS A AMPLIFIER	100 250	−3.0 −3.0	100 100	0.4 0.5	1.8 2.0	1200000 2000000	1225 1300	—	—	—
7D7	TRIODE-HEXODE CONVERTER	8AR	H	6.3	0.43	OSCILLATOR-MIXER	150 250	−3.0 −3.0	VALUES FOR TRIODE UNIT VALUES FOR HEXODE UNIT		3.5	16800	1900	32	CONV. TRANSCOND., 275 MICROMHOS	
7E6	DUPLEX-DIODE TRIODE	8W	H	6.3	0.3	TRIODE UNIT AS CLASS A AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6R7									
7E7	DUPLEX-DIODE PENTODE	8AE	H	6.3	0.3	PENTODE UNIT AS R.F. OR A.F. AMPLIFIER	100 250	−1.0 −3.0	100 100	2.7 1.6	10.0 7.5	150000 700000	1600 1300	—	—	—
7F7	TWIN-TRIODE AMPLIFIER	8AC	H	6.3	0.3	EACH UNIT AS AMPLIFIER	250	−2.0	—	—	2.3	44000	1600	70	—	—
7G7/1232	TRIPLE-GRID AMPLIFIER	8V	H	6.3	0.45	CLASS A AMPLIFIER	250	−2.0	100	2.0	6.0	800000	4500	—	—	—
7H7	R.F. AMPLIFIER PENTODE	8V	H	6.3	0.29	AMPLIFIER	250	−2.5	150	3.5	9.5	800000	3800	—	—	—
7J7	TRIODE-HEXODE CONVERTER	8AR	H	6.3	0.3	CONVERTER	100 250	−3.0 −3.0	100 100	3.1 2.9	1.1 1.3	300000 1500000	260 300	TRIODE PLATE, 250③MAX. VOLTS, 5.4 MA. TRIODE GRID RESISTOR; 50000 Ω; GRID CURRENT, 0.4 MA.		
7L7	TRIPLE-GRID AMPLIFIER	8V	H	6.3	0.3	CLASS A AMPLIFIER	100 250	−1.0 −1.5	100 100	2.4 1.5	5.5 4.5	100000 1000000	3000 3100	—	—	—
7N7	TWIN-TRIODE AMPLIFIER	8AC	H	6.3	0.6	EACH UNIT AS AMPLIFIER	250	−8.0	—	—	9.0	7700	2600	20	—	—
7Q7	PENTAGRID CONVERTER	8AL	H	6.3	0.3	CONVERTER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6SA7									
7R7	DUPLEX-DIODE PENTODE	8AE	H	6.3	0.3	PENTODE UNIT AS AMPLIFIER	250	−1.0	100	2.1	5.7	1000000	3200	—	—	—
7S7	TRIODE-HEXODE CONVERTER	8AR	H	6.3	0.3	OSCILLATOR-MIXER	250 250	TRIODE UNIT HEXODE UNIT	— 100	— 2.2	5.0 1.7	— 2000000	TRIODE GRID CURRENT, 0.4 MA.; RESISTOR, 0.05 MEG. CONV. TRANS., 600 OR 2 MICROMHOS (E_{C3},−2,−22 V.)			
7V7	R.F. AMPLIFIER PENTODE	8V	H	6.3	0.43	AMPLIFIER	300	BIAS RESIS., 160 OHMS	150	3.9	9.6	300000	5800	—	—	—
7Y4	FULL-WAVE RECTIFIER	5AB	H	6.3	0.3	WITH CONDENSER-INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 325 MAX. PEAK INVERSE VOLTS, 1250.				MAX. D.C. OUTPUT MA., 60 MAX. PEAK PLATE MA., 180			MIN. TOTAL EFFECTIVE SUPPLY IMPEDANCE PER PLATE, 150 OHMS		
						WITH CHOKE-INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 450 MAX. PEAK INVERSE VOLTS, 1250.				MAX. D.C. OUTPUT MA., 60 MAX. PEAK PLATE MA., 180			MINIMUM VALUE OF INPUT CHOKE, 10 HENRIES		
7Z4	FULL-WAVE RECTIFIER	5AB	H	6.3	0.86	WITH CONDENSER-INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 325 MAX. PEAK INVERSE VOLTS, 1250				MAX. PEAK PLATE MA., 300			MIN. TOTAL EFFECTIVE SUPPLY IMPEDANCE PER PLATE, 75 OHMS		
						WITH CHOKE-INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 450 MAX. PEAK INVERSE VOLTS, 1250				MAX. PEAK PLATE MA., 300			MINIMUM VALUE OF INPUT CHOKE, 6 HENRIES		
10	POWER AMPLIFIER TRIODE	4D	F	7.5	1.25	CLASS A AMPLIFIER	350 425	−32.0 −40.0	—	—	16.0 18.0	5150 5000	1550 1600	8.0 8.0	11000 10200	0.9 1.6
11 12	DETECTOR ① AMPLIFIER TRIODE	4F 4D	D.C. F	1.1	0.25	CLASS A AMPLIFIER	90 135	−4.5 −10.5	—	—	2.5 3.0	15500 15000	425 440	6.6 6.6	—	—

TYPE	DESIGN	SOCKET CONN.	CATHODE TYPE AND RATING			USED AS	PLATE SUPPLY VOLTS	GRID BIAS (2) VOLTS	SCREEN SUPPLY VOLTS	SCREEN CURRENT MILLIAMP.	PLATE CURRENT MILLIAMP.	A.C. PLATE RESISTANCE OHMS	TRANSCON-DUCTANCE (GRID-PLATE) µMHOS	AMPLI-FICATION FACTOR	LOAD FOR STATED POWER OUTPUT OHMS	POWER OUTPUT WATTS
			C.T.	VOLTS	AMP.											
12A5	POWER AMPLIFIER PENTODE	7F	H	6.3 12.6	0.6 0.3	CLASS A AMPLIFIER	100 180	−15.0 −25.0	100 180	3.0 8.0	6.5 14.0	50000 35000	1700 2400	—	4500 3300	0.8 3.4
12A7	RECTIFIER-PENTODE	7K	H	12.6	0.3	PENTODE UNIT AS CLASS A AMPLIFIER	135	−13.5	135	2.5	9.0	102000	975	—	13500	0.55
						HALF-WAVE RECTIFIER	MAXIMUM A.C. PLATE VOLTAGE .... 125 VOLTS, RMS MAXIMUM D.C. OUTPUT CURRENT .... 30 MILLIAMPERES									
12A8 (G) (GT)	PENTAGRID CONVERTER (8)	8A	H	12.6	0.15	CONVERTER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6A8-GT									
12AH7-GT	TWIN TRIODE AMPLIFIER	8BE	H	12.6	0.15	AMPLIFIER	250	−9	—	—	12.0	6600	2400	16	—	—
12B7	TRIPLE-GRID SUPER-CONTROL AMPLIFIER	8V	H	12.6	0.15	CLASS A AMPLIFIER	100 250	−3.0 −3.0	100 100	2.6 2.4	8.9 9.2	250000 800000	1900 2000	—	—	—
12B8-GT	TRIODE PENTODE	8T	H	12.6	0.3	TRIODE UNIT AS CLASS A AMPLIFIER	90 100	0 −1.0	—	—	2.8 0.6	37000 73000	2400 1500	90 110	—	—
						PENTODE UNIT AS CLASS A AMPLIFIER	90 100	−3.0 −3.0	90 100	2.0 2.0	7.0 8.0	170000 200000	1800 2100	360 360	—	—
12C8	DUPLEX-DIODE PENTODE	8E	H	12.6	0.15	PENTODE UNIT AS R.F. AMPLIFIER	250	−3.0	125	2.3	10.0	600000	1325	—	—	—
						PENTODE UNIT AS A.F. AMPLIFIER	90 (18) 300 (18)	CATHODE BIAS, 3500 OHMS. SCREEN RESISTOR= 1.1 MEG. CATHODE BIAS, 1600 OHMS. SCREEN RESISTOR= 1.2 MEG.					GRID RESISTOR, (10) 0.5 MEGOHM	GAIN PER STAGE = 55 GAIN PER STAGE = 79		
12E5-GT	AMPLIFIER TRIODE	6Q	H	12.6	0.15	CLASS A AMPLIFIER	250	−13.5	—	—	5.0	9500	1450	13.8	—	—
12F5-GT	HIGH-MU TRIODE	5M	H	12.6	0.15	CLASS A AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6SF5.									
12G7-G (GT)	DUPLEX-DIODE HIGH-MU TRIODE	7V	H	12.6	0.15	TRIODE UNIT AS CLASS A AMPLIFIER	250	−3.0	—	—	—	58000	1200	70	—	—
12H6	TWIN DIODE	7Q	H	12.6	0.15	DETECTOR RECTIFIER	FOR OTHER CHARACTERISTICS REFER TO TYPE 6H6									
12J5-GT	DETECTOR AMPLIFIER TRIODE	6Q	H	12.6	0.15	AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6J5.									
12J7-GT	TRIPLE-GRID DETECTOR AMPLIFIER	7R	H	12.6	0.15	AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6J7.									
12K7-G (GT)	TRIPLE-GRID SUPER-CONTROL AMPLIFIER	7R	H	12.6	0.15	AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6K7-GT									
12K8 (GT)	TRIODE-HEXODE CONVERTER	8K	H	12.6	0.15	OSCILLATOR MIXER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6K8.									
12Q7-G (GT)	DUPLEX-DIODE HIGH-MU TRIODE	7V	H	12.6	0.15	TRIODE UNIT AS AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6Q7-GT									

12SA7(GT)	PENTAGRID CONVERTER (26)	8R 8AD	H	12.6	0.15	MIXER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6SA7									
12SC7	TWIN-TRIODE AMPLIFIER	8S	H	12.6	0.15	AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6SC7									
12SF5(GT)	HIGH-MU TRIODE	6AB	H	12.6	0.15	AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6SF5									
12SF7	DIODE-SUPER-CONTROL AMPLIFIER PENTODE	7AZ	H	12.6	0.15	PENTODE UNIT AS CLASS A AMPLIFIER	FOR OTHER CHARACTERISTICS REFER TO TYPE 6SF7									
12SG7	H.F. AMPLIFIER PENTODE	8BC	H	12.6	0.15	CLASS A AMPLIFIER	FOR OTHER CHARACTERISTICS REFER TO TYPE 6SG7									
12SJ7(GT)	TRIPLE-GRID DETECTOR AMPLIFIER	8N	H	12.6	0.15	AMPLIFIER	FOR OTHER CHARACTERISTICS REFER TO TYPE 6SJ7 AND TO TYPE 6SJ7-GT									
12SK7(GT)	TRIPLE-GRID SUPER-CONTROL AMPLIFIER	8N	H	12.6	0.15	AMPLIFIER	FOR OTHER CHARACTERISTICS REFER TO TYPE 6SK7 AND TO TYPE 6SK7-GT									
12SL7-GT	TWIN-TRIODE AMPLIFIER	8BD	H	12.6	0.15	CLASS A AMPLIFIER (39)	FOR OTHER CHARACTERISTICS REFER TO TYPE 6SL7-GT									
12SN7-GT	TWIN-TRIODE AMPLIFIER	8BD	H	12.6	0.3	CLASS A AMPLIFIER (39)	FOR OTHER CHARACTERISTICS REFER TO TYPE 6SN7-GT									
12SQ7-GT/G	DUPLEX-DIODE HIGH-MU TRIODE	8Q	H	12.6	0.15	TRIODE UNIT AS AMPLIFIER	FOR OTHER CHARACTERISTICS REFER TO TYPE 6SQ7									
12SR7	DUPLEX-DIODE TRIODE	8Q	H	12.6	0.15	TRIODE UNIT AS AMPLIFIER	FOR OTHER CHARACTERISTICS REFER TO TYPE 6SR7									
12Z3	HALF-WAVE RECTIFIER	4G	H	12.6	0.3	WITH CONDENSER-INPUT FILTER	MAX. A.C. PLATE VOLTS (RMS), 235 MAX. D.C. OUTPUT MA., 55				MIN. TOTAL EFFECTIVE PLATE-SUPPLY IMPEDANCE: UP TO 117 VOLTS, 0 OHMS; AT 150 VOLTS, 30 OHMS; AT 235 VOLTS, 75 OHMS.					
14A4	DETECTOR AMPLIFIER TRIODE	5AC	H	12.6	0.15	CLASS A AMPLIFIER	FOR OTHER CHARACTERISTICS REFER TO TYPE 7A4									
14A5	BEAM POWER AMPLIFIER	6AA	H	12.6	0.15	CLASS A AMPLIFIER	250	−12.5	250	3.5	30.0	50000	3000	—	7500	2.5
14A7/12B7	TRIPLE-GRID SUPER-CONTROL AMPLIFIER	8V	H	12.6	0.15	CLASS A AMPLIFIER	100 250	−1.0 −3.0	100 100	4.0 2.6	13.0 9.2	120000 800000	2350 2000	—	—	—
14B6	DUPLEX-DIODE HIGH-MU TRIODE	8W	H	12.6	0.15	TRIODE UNIT AS CLASS A AMPLIFIER	250	−2.0	—	—	0.9	91000	1100	100	—	—
14B8	PENTAGRID CONVERTER	8X	H	12.6	0.15	CONVERTER	FOR OTHER CHARACTERISTICS REFER TO TYPE 7B8-LM									
14C5	BEAM POWER AMPLIFIER	6AA	H	12.6	0.225	CLASS A AMPLIFIER	FOR OTHER CHARACTERISTICS REFER TO TYPE 6V6									
14C7	R.F. AMPLIFIER PENTODE	8V	H	12.6	0.15	AMPLIFIER	250	−3.0	100	0.7	2.2	1000000	1575	—	—	—
14E6	DUPLEX-DIODE TRIODE	8W	H	12.6	0.15	TRIODE UNIT AS CLASS A AMPLIFIER	FOR OTHER CHARACTERISTICS REFER TO TYPE 6R7									
14E7	DUPLEX-DIODE PENTODE	8AE	H	12.6	0.15	PENTODE UNIT AS R.F. OR A.F. AMPLIFIER	FOR OTHER CHARACTERISTICS REFER TO TYPE 7E7									
14F7	TWIN-TRIODE AMPLIFIER	8AC	H	12.6	0.15	EACH UNIT AS AMPLIFIER	FOR OTHER CHARACTERISTICS REFER TO TYPE 7F7									
14H7	R.F. AMPLIFIER PENTODE	8V	H	12.6	0.14	AMPLIFIER	250	−2.5	150	3.5	9.5	800000	3800	—	—	—
14J7	TRIODE-HEXODE CONVERTER	8AR	H	12.6	0.15	CONVERTER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 7J7									
14N7	TWIN-TRIODE AMPLIFIER	8AC	H	12.6	0.3	EACH UNIT AS AMPLIFIER	FOR OTHER CHARACTERISTICS REFER TO TYPE 7N7									
14Q7	PENTAGRID CONVERTER	8AL	H	12.6	0.15	CONVERTER	250	−2.0	100	8.5	3.5	1000000	450	OSCILLATOR GRID (#1) RESISTOR, 20000 OHMS. OSCILLATOR GRID CURRENT, 0.5 MA.		

TYPE	DESIGN	SOCKET CONN.	CATHODE TYPE AND RATING C.T.	VOLTS	AMP.	USED AS	PLATE SUPPLY VOLTS	GRID BIAS ② VOLTS	SCREEN SUPPLY VOLTS	SCREEN CURRENT MILLIAMP.	PLATE CURRENT MILLIAMP.	A.C. PLATE RESISTANCE OHMS	TRANSCON-DUCTANCE (GRID-PLATE) µMHOS	AMPLI-FICATION FACTOR	LOAD FOR STATED POWER OUTPUT OHMS	POWER OUTPUT WATTS
14R7	DUPLEX-DIODE PENTODE	8AE	H	12.6	0.15	PENTODE UNIT AS AMPLIFIER	FOR OTHER CHARACTERISTICS REFER TO TYPE 7R7									
14Y4	FULL-WAVE RECTIFIER	5AB	H	12.6	0.3	WITH CONDENSER INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 325 MAX. PEAK INVERSE VOLTS, 1250				MAX. D.C. OUTPUT MA., 70 MAX. PEAK PLATE MA., 210			MIN. TOTAL EFFECTIVE SUPPLY IMPEDANCE, 150 OHMS		
						WITH CHOKE-INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 450 MAX. PEAK INVERSE VOLTS, 1250				MAX. D.C. OUTPUT MA., 70 MAX. PEAK PLATE MA., 210			—		
15	R.F. AMPLIFIER PENTODE	5F	D.C. H	2.0	0.22	CLASS A AMPLIFIER	67.5 135	−1.5 −1.5	67.5 67.5	0.3 0.3	1.85 1.85	630000 800000	710 750	450 600	—	—
18	POWER AMPLIFIER PENTODE	6B	H	14.0	0.3	CLASS A AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6F6									
19	TWIN-TRIODE AMPLIFIER	6C	D.C. F	2.0	0.26	AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 1J6-G									
20	POWER AMPLIFIER TRIODE	4D	D.C. F	3.3	0.132	CLASS A AMPLIFIER	90 135	−16.5 −22.5	—	—	3.0 6.5	8000 6300	415 525	3.3 3.3	9600 6500	0.045 0.110
20J8	TRIODE-HEPTODE CONVERTER	8H	H	20.0	0.15		FOR OTHER CHARACTERISTICS, REFER TO TYPE 6J8-G									
21A7	TRIODE-HEXODE CONVERTER	8AR	H	21.0	0.16	OSCILLATOR-MIXER	150 250	−3.0 −3.0	VALUES FOR TRIODE UNIT VALUES FOR HEXODE UNIT		3.5 —	16800 1500000	1900 —	CONVERSION TRANSCONDUCTANCE 275 MICROMHOS		
22	R.F AMPLIFIER TETRODE	4K	D.C. F	3.3	0.132	SCREEN-GRID R.F AMPLIFIER	135 135	−1.5 −1.5	45 67 5	0.6 ㉗ 1.3 ㉗	1.7 3.7	725000 325000	375 500	270 160	—	—
24-A	R.F. AMPLIFIER PENTODE	5E	H	2.5	1 75	SCREEN-GRID R.F AMPLIFIER	180 250	−3.0 −3.0	90 90	1 7 ㉗ 1.7 ㉗	4.0 4.0	400000 600000	1000 1050	400 630	—	—
						BIAS DETECTOR	250 ⑯	{−5.0 APPROX.}	20 TO 45	PLATE CURRENT ADJUSTED TO 0.1 MA. WITH NO SIGNAL						
25A6-GT/G	POWER AMPLIFIER PENTODE	7S	H	25.0	0.3	CLASS A AMPLIFIER	95 160	−15.0 −18.0	95 120	4.0 6.5	20.0 33.0	45000 42000	2000 2375	—	4500 5000	0.9 2.2
25A7-GT/G	RECTIFIER-PENTODE	8F	H	25.0	0.3	PENTODE UNIT AS CLASS A AMPLIFIER	100	−15.0	100	4.0	20.5	50000	1800	—	4500	0.77
						HALF-WAVE RECTIFIER		MAX. A.C PLATE VOLTAGE . . . MAX D.C OUTPUT CURRENT. . .			117 VOLTS, RMS 75 MILLIAMPERES					
25AC5-GT/G	HIGH-MU POWER AMPLIFIER TRIODE	6Q	H	25 0	0.3	CLASS B AMPLIFIER	180	0	—	—	4.0 ⑬	—	—	—	4800	6.0
						DYNAMIC-COUPLED AMP. WITH TYPE 6AE5-GT DRIVER	110	BIAS FOR BOTH 25AC5-GT AND 6AE5-GT DEVELOPED IN CIRCUIT AVERAGE PLATE CURRENT OF DRIVER = 7 MILLIAMPERES AVERAGE PLATE CURRENT OF 25AC5-GT = 45 MILLIAMPERES							2000	2.0
25B5	DIRECT-COUPLED TRIODES	6D	H	25.0	0.3	AMPLIFIER	IN. PLATE 100	0	OUT PLATE 180	OUT. PLATE 46	IN. PLATE 5.8	—	—	—	4000	3.8
25B6-G	POWER AMPLIFIER PENTODE	7S	H	25.0	0.3	CLASS A AMPLIFIER	105 200	−16.0 −23.0	105 135	2.0 1.8	48.0 62.0	15500 18000	4800 5000	—	1700 2500	2.4 7.1
25B8-GT	TRIODE-PENTODE	8T	H	25 0	0 15	PENTODE UNIT AS AMPLIFIER	100	−3.0	100	2.0	7.6	185000	2000	—	—	—
						TRIODE UNIT AS AMPLIFIER	100	−1.0	—	—	0.6	75000	1500	112	—	—

25C6-G	BEAM POWER AMPLIFIER	7AC	H	25.0	0.3	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6Y6-G.										
25L6-GT/G	BEAM POWER AMPLIFIER	7AC	H	25.0	0.3	AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 50L6-GT.									
25N6 (G)	DIRECT-COUPLED TRIODES	7W	H	25.0	0.3	AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 25B5									
25S/1B5	DUPLEX-DIODE TRIODE	REFER TO TYPE 1B5 DATA														
25X6-GT	FULL-WAVE RECTIFIER	7Q	H	25.0	0.15	RECTIFIER	MAX. A.C VOLTS PER PLATE (RMS), 125 MAX. D.C. OUTPUT CURRENT, 75 MA.									
25Y4-GT	HALF-WAVE RECTIFIER	5AA	H	25.0	0.15	RECTIFIER	MAX. A.C. PLATE VOLTAGE . . . .125 VOLTS, RMS MAX. D.C. OUTPUT CURRENT. . . . . . .75 MILLIAMPERES									
25Y5	RECTIFIER-DOUBLER	6E	H	25.0	0.3	RECTIFIER	MAX. A.C. VOLTS PER PLATE (RMS), 235 MAX. D.C. OUTPUT CURRENT. . . . . . . .75 MILLIAMPERES } HALF-WAVE OPERATION									
25Z4-GT	HALF-WAVE RECTIFIER	5AA	H	25.0	0.3	RECTIFIER	MAX. A.C. PLATE VOLTAGE . . . . . . 125 VOLTS, RMS MAX. D.C. OUTPUT CURRENT . . . . . 125 MILLIAMPERES									
25Z5	RECTIFIER-DOUBLER	6E	H	25.0	0.3	RECTIFIER-DOUBLER	FOR OTHER RATINGS, REFER TO TYPE 25Z6.									
25Z6-GT/G	RECTIFIER-DOUBLER	7Q	H	25.0	0.3	VOLTAGE DOUBLER	MAX. A.C. VOLTS PER PLATE (RMS), 117 MAX. D.C. OUTPUT MA., 75				MIN. TOTAL EFFECTIVE PLATE-SUPPLY IMPEDANCE: HALF-WAVE, 30 OHMS; FULL-WAVE, 0 OHMS.					
						HALF-WAVE RECTIFIER	MAX. PLATE VOLTAGE (RMS), 235 MAX. D.C. OUTPUT MA. PER PLATE, 75				MIN. TOTAL EFFECTIVE SUPPLY IMPEDANCE PER PLATE: UP TO 117 VOLTS, 0 OHMS; AT 150 VOLTS, 40 OHMS; AT 235 VOLTS, 100 OHMS.					
26	AMPLIFIER-TRIODE	4D	F	1.5	1.05	CLASS A AMPLIFIER	90 180	−7.0 −14.5	——	——	2.9 6.2	8900 7300	935 1150	8.3 8.3	——	——
27	DETECTOR (1) AMPLIFIER TRIODE	5A	H	2.5	1.75	CLASS A AMPLIFIER	135 250	−9.0 −21.0	——	——	4.5 5.2	9000 9250	1000 975	9.0 9.0	——	——
						BIAS DETECTOR	250	{−30.0 APPROX.}	——	——	PLATE CURRENT ADJUSTED TO 0.2 MILLIAMPERE WITH NO SIGNAL.					
30	DETECTOR (1) AMPLIFIER TRIODE	4D	D.C. F	2.0	0.06	AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 1H4-G.									
31	POWER AMPLIFIER TRIODE	4D	D.C. F	2.0	0.13	CLASS A AMPLIFIER	135 180	−22.5 −30.0	——	——	8.0 12.3	4100 3600	925 1050	3.8 3.8	7000 5700	0.185 0.375
32	R.F. AMPLIFIER TETRODE	4K	D.C. F	2.0	0.06	SCREEN-GRID R.F. AMPLIFIER	135 180	−3.0 −3.0	67.5 67.5	0.4 (27) 0.4 (27)	1.7 1.7	950000 1200000	640 650	——	——	——
						BIAS DETECTOR	180 (17)	{−6.0 APPROX.}	67.5	——	PLATE CURRENT ADJUSTED TO 0.2 MILLIAMPERE WITH NO SIGNAL.					
32L7-GT	RECTIFIER-BEAM POWER AMPLIFIER	8Z	H	32.5	0.3	H.W. RECTIFIER	MAX. A.C. PLATE VOLTS (RMS), 125. MAX. D.C. OUTPUT CURRENT, 60 MILLIAMPERES									
						CLASS A AMPLIFIER	110	−7.5	110	3.0	40.0	15000	6000	——	2500	1.5
33	POWER AMPLIFIER PENTODE	5K	D.C. F	2.0	0.26	CLASS A AMPLIFIER	180	−18.0	180	5.0	22.0	55000	1700	——	6000	1.4
34	SUPER-CONTROL R.F. AMPLIFIER PENTODE	4M	D.C. F	2.0	0.06	SCREEN-GRID R.F. AMPLIFIER	135 180	{−3.0 MIN.}	67.5 67.5	1.0 1.0	2.8 2.8	600000 1000000	360 620	——	——	——
35/51	SUPER-CONTROL R.F. AMPLIFIER TETRODE	5E	H	2.5	1.75	SCREEN-GRID R.F. AMPLIFIER	180 250	{−3.0 MIN.}	90.0 90.0	2.5 (27) 2.5 (27)	6.3 6.5	300000 400000	1020 1050	305 420	——	——
35A5-LT	BEAM POWER AMPLIFIER	6AT	H	35.0	0.15	SINGLE-TUBE CLASS A AMPLIFIER	110	−7.5	110	3.0	40.0	14000	5800	——	2500	1.5
35L6-GT/G	BEAM POWER AMPLIFIER	7AC	H	35.0	0.15	SINGLE-TUBE CLASS A AMPLIFIER	110	−7.5	110	3.0	40.0	13800	5800	——	2500	1.5

TYPE	DESIGN	SOCKET CONN.	CATHODE TYPE AND RATING C.T.	VOLTS	AMP.	USED AS	PLATE SUPPLY VOLTS	GRID BIAS (2) VOLTS	SCREEN SUPPLY VOLTS	SCREEN CURRENT MILLIAMP.	PLATE CURRENT MILLIAMP.	A.C. PLATE RESISTANCE OHMS	TRANSCONDUCTANCE (GRID-PLATE) µMHOS	AMPLIFICATION FACTOR	LOAD FOR STATED POWER OUTPUT OHMS	POWER OUTPUT WATTS
35Z3-LT	HALF-WAVE RECTIFIER	4Z	H	35.0	0.15	WITH CONDENSER-INPUT FILTER	MAX. A.C. PLATE VOLTS (RMS), 235 MAX. D.C. OUTPUT MA., 100				MIN. TOTAL EFFECTIVE PLATE-SUPPLY IMPEDANCE: UP TO 117 VOLTS, 15 OHMS; AT 150 VOLTS, 40 OHMS; AT 235 VOLTS, 100 OHMS.					
35Z4-GT	HALF-WAVE RECTIFIER	5AA	H	35.0	0.15	WITH CONDENSER-INPUT FILTER	MAX. A.C. PLATE VOLTS (RMS), 250 (22) MAX. PEAK INVERSE VOLTS, 720				MAX. D.C. OUTPUT MA., 100 MAX. PEAK PLATE MA., 600					
35Z5-GT/G	HALF-WAVE RECTIFIER *Heater Tap for Pilot*	6AD	H	35.0	0.15	WITH CONDENSER-INPUT FILTER	MAX. A.C. PLATE VOLTS (RMS), 235, MIN. TOTAL EFFECTIVE PLATE-SUPPLY IMPEDANCE: UP TO 117 VOLTS, 0 OHMS, AT 235 VOLTS, 45 OHMS. MAX. D.C. OUTPUT MA. WITH PILOT AND NO SHUNT RESISTOR, 60; WITH PILOT AND SHUNT RESISTOR, 100; WITHOUT PILOT, 100									
35Z6-G	RECTIFIER-DOUBLER	7Q	H	35.0	0.3	RECTIFIER-DOUBLER	MAX. A.C. VOLTS PER PLATE (RMS), 235 MAX. D.C. OUTPUT MA., 110									
36	R.F. AMPLIFIER TETRODE	5E	H	6.3	0.3	SCREEN-GRID R.F. AMPLIFIER	100 250	−1.5 −3.0	55 90	— 1.7 (27)	1.8 3.2	550000 550000	850 1080	470 595	—	—
						BIAS DETECTOR	100 (16) 250 (16)	−5.0 −8.0	55 90	—	GRID-BIAS VALUES ARE APPROXIMATE. PLATE CURRENT ADJUSTED TO 0.1 MILLIAMPERE WITH NO SIGNAL.					
37	DETECTOR (1) AMPLIFIER TRIODE	5A	H	6.3	0.3	CLASS A AMPLIFIER	90 250	−6.0 −18.0	—	—	2.5 7.5	11500 8400	800 1100	9.2 9.2	—	—
						BIAS DETECTOR	90 250	−10.0 −28.0	—	—	GRID BIAS VALUES ARE APPROXIMATE. PLATE CURRENT ADJUSTED TO 0.2 MILLIAMPERE WITH NO SIGNAL.					
38	POWER AMPLIFIER PENTODE	5F	H	6.3	0.3	CLASS A AMPLIFIER	100 250	−9.0 −25.0	100 250	1.2 3.8	7.0 22.0	140000 100000	875 1200	—	15000 10000	0.27 2.50
39/44	SUPER-CONTROL R.F. AMPLIFIER PENTODE	5F	H	6.3	0.3	CLASS A AMPLIFIER	90 250	{−3.0 MIN.}	90 90	1.6 1.4	5.6 5.8	375000 1000000	960 1050	360 1050	—	—
40	VOLTAGE AMPLIFIER TRIODE	4D	D.C. F	5.0	0.25	CLASS A AMPLIFIER	135 (18) 180 (18)	−1.5 −3.0	—	—	0.2 0.2	150000 150000	200 200	30 30	—	—
40Z5/45Z5-GT	HALF-WAVE RECTIFIER	REFER TO TYPE 45Z5-GT DATA														
41	POWER AMPLIFIER PENTODE	6B	H	6.3	0.4	AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6K6-G.									
42	POWER AMPLIFIER PENTODE	6B	H	6.3	0.7	AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6F6.									
43	POWER AMPLIFIER PENTODE	6B	H	25.0	0.3	AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 25A6.									
44	SUPER-CONTROL R.F. AMPLIFIER PENTODE	REFER TO TYPE 39/44 DATA														
45	POWER AMPLIFIER TRIODE	4D	F	2.5	1.5	CLASS A AMPLIFIER	180 275	−31.5 −56.0	—	—	31.0 36.0	1650 1700	2125 2050	3.5 3.5	2700 4600	0.82 2.0
						PUSH-PULL CLASS AB_2 AMPLIFIER	275 275	CATHODE BIAS, 775 OHMS (13) −68.0 VOLTS, FIXED BIAS			36.0 (13) 28.0 (13)	—	—	—	5060 3200	12.0 (12) 18.0 (12)
45Z3	HALF-WAVE RECTIFIER	5AM	H	45.0	0.075	WITH CONDENSER-INPUT FILTER	MAX. A.C. PLATE VOLTAGE (RMS), 117 VOLTS MAX. PEAK INVERSE VOLTAGE, 350 VOLTS				MAX. D.C. OUTPUT MA., 65 MAX. PEAK PLATE MA., 390				MIN. TOTAL EFFECTIVE SUPPLY IMPEDANCE, 15 OHMS	

45Z5-GT	HALF-WAVE RECTIFIER *Heater Tap for Pilot*	6AD	H	45.0	0.15	WITHOUT PILOT	MAX. A.C. PLATE VOLTS (RMS), 250 (22) MAX. PEAK PLATE MA., 600 MAX. D.C. OUTPUT MA., 100									
						WITH PILOT	MAX. A.C. PLATE VOLTS RMS 250 (22) MAX. D.C. OUTPUT MA., 60									
46	DUAL-GRID POWER AMPLIFIER	5C	F	2.5	1.75	CLASS A AMPLIFIER (21)	250	−33.0	—	—	22.0	2380	2350	5.6	6400	1.25
						CLASS B AMPLIFIER (29)	300 400	0 0	—	—	8.0 (13) 12.0 (13)	—	—	—	5200 5800	16.0 (12) 20.0 (12)
47	POWER AMPLIFIER PENTODE	5B	F	2.5	1.75	CLASS A AMPLIFIER	250	−16.5	250	6.0	31.0	60000	2500	150	7000	2.7
48	POWER AMPLIFIER TETRODE	6A	D.C. H	30.0	0.4	TETRODE CLASS A AMPLIFIER	96 125	−19.0 −20.0	96 100	9.0 9.5	52.0 56.0	—	3800 3900	—	1500 1500	2.0 2.5
						TETRODE PUSH-PULL CLASS A AMPLIFIER	125	−20.0	100	—	100 (13)	—	—	—	3000	5.0 (12)
49	DUAL-GRID POWER AMPLIFIER	5C	D.C. F	2.0	0.12	CLASS A AMPLIFIER (21)	135	−20.0	—	—	6.0	4175	1125	4.7	11000	0.17
						CLASS B AMPLIFIER (29)	180	0	—	—	4.0 (13)	—	—	—	12000	3.5 (12)
50	POWER AMPLIFIER TRIODE	4D	F	7.5	1.25	CLASS A AMPLIFIER	300 400 450	−54.0 −70.0 −84.0	—	—	35.0 55.0 55.0	2000 1800 1800	1900 2100 2100	3.8 3.8 3.8	4600 3670 4350	1.6 3.4 4.6
50A5	BEAM POWER AMPLIFIER	6AA	H	50.0	0.15	CLASS A AMPLIFIER	200	−8.0	110	1.5	50.0	35000	8250	—	3000	4.7
50C6-G	BEAM POWER AMPLIFIER	7AC	H	50.0	0.15	AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6Y6-G.									
50L6-GT	BEAM POWER AMPLIFIER	7AC	H	50.0	0.15	SINGLE-TUBE CLASS A AMPLIFIER	110 110	−7.5 −7.5	110 110	4.0 4.0	49.0 49.0	10000 10000	8200 8200	—	1500 2000	2.1 2.2
50Y6-GT	RECTIFIER DOUBLER	7Q	H	50.0	0.3	VOLTAGE DOUBLER	MAX. A.C. VOLTS PER PLATE (RMS), 117 MAX. D.C. OUTPUT MA., 75				MIN. TOTAL EFFECTIVE PLATE-SUPPLY IMPEDANCE: HALF-WAVE, 30 OHMS; FULL WAVE, 15 OHMS.					
						HALF-WAVE RECTIFIER	MAX. PLATE VOLTAGE (RMS), 235 MAX. D.C. OUTPUT MA. PER PLATE; 75				MIN. TOTAL EFFECTIVE SUPPLY IMPEDANCE PER PLATE: UP TO 117 VOLTS, 15 OHMS; AT 150 VOLTS, 40 OHMS; AT 235 VOLTS, 100 OHMS.					
50Z6-G	FULL-WAVE RECTIFIER	7Q	H	50.0	0.3	RECTIFIER	MAX. A.C. PLATE VOLTS (RMS), 250 MAX. D.C. OUTPUT MA., 250									
50Z7-G	RECTIFIER DOUBLER	8AN	H	50.0	0.15	RECTIFIER DOUBLER	MAX. A.C. PLATE VOLTS (RMS), 117. MAX. D.C. OUTPUT CURRENT WHEN USED WITH 2.9-VOLT 0.17-AMP. PANEL LAMP, 65 MILLIAMPERES.									
51	SUPER-CONTROL R.F. AMPLIFIER TETRODE	REFER TO TYPE 35/51 DATA														
52	DUAL-GRID POWER AMPLIFIER	5C	F	6.3	0.3	CLASS A AMPLIFIER (21)	110	0	—	—	43.0	1750	3000	5.2	2000	1.5
						CLASS B AMPLIFIER (29)	180	0	—	—	3.0 (13)	—	—	—	10000	5.0 (12)
53	TWIN-TRIODE AMPLIFIER	7B	H	2.5	2.0	AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6N7.									
55	DUPLEX-DIODE TRIODE	6G	H	2.5	1.0	TRIODE UNIT AS AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 85.									
56	DETECTOR AMPLIFIER TRIODE (1)	5A	H	2.5	1.0	AMPLIFIER DETECTOR	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6P5-G.									
57	TRIPLE-GRID DETECTOR AMPLIFIER	6F	H	2.5	1.0	AMPLIFIER DETECTOR	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6J7.									
58	TRIPLE-GRID SUPER-CONTROL AMPLIFIER	6F	H	2.5	1.0	AMPLIFIER MIXER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6U7-G.									

TYPE	DESIGN	SOCKET CONN.	CATHODE TYPE AND RATING C.T.	VOLTS	AMP.	USED AS	PLATE SUPPLY VOLTS	GRID BIAS ② VOLTS	SCREEN SUPPLY VOLTS	SCREEN CURRENT MILLIAMP.	PLATE CURRENT MILLIAMP.	A.C. PLATE RESISTANCE OHMS	TRANSCONDUCTANCE (GRID-PLATE) µMHOS	AMPLIFICATION FACTOR	LOAD FOR STATED POWER OUTPUT OHMS	POWER OUTPUT WATTS
59	TRIPLE-GRID POWER AMPLIFIER	7A	H	2.5	2.0	TRIODE ⑥ CLASS A AMPLIFIER	250	−28.0	—	—	26.0	2300	2600	6.0	5000	1.25
						PENTODE ⑤ CLASS A AMPLIFIER	250	−18.0	250	9.0	35.0	40000	2500	—	6000	3.0
						TRIODE ⑦ CLASS B AMPLIFIER	300 400	0 0	—	—	20.0 ⑬ 26.0 ⑬	—	—	—	4600 6000	15.0 ⑫ 20.0 ⑫
70A7-GT	RECTIFIER-BEAM POWER AMPLIFIER *Heater Tap for Pilot*	8E	H	70	0.3	HALF-WAVE RECTIFIER	MAX. A.C. PLATE VOLTS (RMS), 125. MAX. D.C. OUTPUT CURRENT, 60 MA.									
						CLASS A AMPLIFIER	110	−7.5	110	3.0	40.0		5800	80	2500	1.5
70L7-GT	RECTIFIER-BEAM POWER AMPLIFIER	8AA	H	70	0.15	HALF-WAVE RECTIFIER	MAX. A.C. PLATE VOLTS (RMS), 117. MAX. D.C. OUTPUT CURRENT, 70 MA.									
						CLASS A AMPLIFIER	110	−7.5	110	3.0	43.0	15000	7500	—	2000	1.8
71-A	POWER AMPLIFIER TRIODE	4D	F	5.0	0.25	CLASS A AMPLIFIER	90 180	−16.5 −40.5	—	—	10.0 20.0	2170 1750	1400 1700	3.0 3.0	3000 4800	0.125 0.790
75	DUPLEX-DIODE HIGH-MU TRIODE	6G	H	6.3	0.3	AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6SQ7.									
76	DETECTOR AMPLIFIER TRIODE ①	5A	H	6.3	0.3	AMPLIFIER DETECTOR	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6P5-G.									
77	TRIPLE-GRID DETECTOR AMPLIFIER	6F	H	6.3	0.3	CLASS A AMPLIFIER	100 250	−1.5 −3.0	60 100	0.4 0.5	1.7 2.3	600000 1000000+ ㉘	1100 1250	—	—	—
						BIAS DETECTOR	250	−1.95	50	CATHODE CURRENT 0.65 MA.		—	PLATE RESISTOR, 250000 OHMS. GRID RESISTOR ⑩ 250000 OHMS.			
78	TRIPLE-GRID SUPER-CONTROL AMPLIFIER	6F	H	6.3	0.3	AMPLIFIER MIXER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6K7.									
79	TWIN-TRIODE AMPLIFIER	6H	H	6.3	0.6	CLASS B AMPLIFIER	180 250	0 0	—	—	7.6 10.6	POWER OUTPUT IS FOR ONE TUBE AT STATED PLATE-TO-PLATE LOAD			7000 14000	5.5 8.0
80	FULL-WAVE RECTIFIER	4C	F	5.0	2.0	RECTIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 5Y3-G.									
81	HALF-WAVE RECTIFIER	4B	F	7.5	1.25	WITH CONDENSER-INPUT FILTER	MAX. A.C. PLATE VOLTS (RMS), 700. MAX. PEAK INVERSE VOLTS, 2000				MAX. D.C. OUTPUT MA., 85 MAX. PEAK PLATE MA., 500					
82	FULL-WAVE RECTIFIER ④	4C	F	2.5	3.0	WITH CONDENSER-INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 450 MAX. PEAK INVERSE VOLTS, 1550				MAX. D.C. OUTPUT MA., 115 MAX. PEAK PLATE MA., 345			MIN. TOTAL EFFECTIVE SUPPLY IMPEDANCE PER PLATE, 50 OHMS		
						WITH CHOKE-INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 550 MAX. PEAK INVERSE VOLTS, 1550				MAX. D.C. OUTPUT MA., 115 MAX. PEAK PLATE MA., 345			MINIMUM VALUE OF INPUT CHOKE, 6 HENRIES		
83	FULL-WAVE RECTIFIER ④	4C	F	5.0	3.0	WITH CONDENSER-INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 450 MAX. PEAK INVERSE VOLTS, 1550				MAX. D.C. OUTPUT MA., 225 MAX. PEAK PLATE MA., 675			MIN. TOTAL EFFECTIVE SUPPLY IMPEDANCE PER PLATE, 50 OHMS		
						WITH CHOKE-INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 550 MAX. PEAK INVERSE VOLTS, 1550				MAX. D.C. OUTPUT MA., 225 MAX. PEAK PLATE MA., 675			MINIMUM VALUE OF INPUT CHOKE, 3 HENRIES		
83-V	FULL-WAVE RECTIFIER	4L	H	5.0	2.0	FOR OTHER RATINGS, REFER TO TYPE 5V4-G.										

84/6Z4	FULL-WAVE RECTIFIER	5D	H	6.3	0.5	WITH CONDENSER-INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 325 MAX. PEAK INVERSE VOLTS, 1250				MAX. D.C. OUTPUT MA., 60 MAX. PEAK PLATE MA., 180			MIN. TOTAL EFFECTIVE SUPPLY IMPEDANCE PER PLATE, 65 OHMS		
						WITH CHOKE-INPUT FILTER	MAX. A.C. VOLTS PER PLATE (RMS), 450 MAX. PEAK INVERSE VOLTS, 1250				MAX. D.C. OUTPUT MA., 60 MAX. PEAK PLATE MA., 180			MINIMUM VALUE OF INPUT CHOKE, 10 HENRIES		
85	DUPLEX-DIODE TRIODE	6G	H	6.3	0.3	TRIODE UNIT AS CLASS A AMPLIFIER	135 250	−10.5 −20.0	——	——	3.7 8.0	11000 7500	750 1100	8.3 8.3	25000 20000	0.075 0.350
89	TRIPLE-GRID POWER AMPLIFIER	6F	H	6.3	0.4	AS TRIODE (6) CLASS A AMPLIFIER	160 250	−20.0 −31.0	——	——	17.0 32.0	3300 2600	1425 1800	4.7 4.7	7000 5500	0.30 0.90
						AS PENTODE (5) CLASS A AMPLIFIER	100 250	−10.0 −25.0	100 250	1.6 5.5	9.5 32.0	104000 70000	1200 1800	——	10700 6750	0.33 3.40
						AS TRIODE (7) CLASS B AMPLIFIER	180	0	——	——	6.0 (13)	——	——	——	13600 9400	2.50 (12) 3.50 (12)
V-99 X-99	DETECTOR (1) AMPLIFIER TRIODE	4E 4D	D.C. F	3.3	0.063	CLASS A AMPLIFIER	90	−4.5	——	——	2.5	15500	425	6.6	——	——
112-A	DETECTOR (1) AMPLIFIER TRIODE	4D	D.C. F	5.0	0.25	CLASS A AMPLIFIER	90 180	−4.5 −13.5	——	——	5.0 7.7	5400 4700	1575 1800	8.5 8.5	5000 10650	0.035 0.285
117L7-GT/ 117M7-GT	RECTIFIER-BEAM POWER AMPLIFIER *Heater Tap for Pilot*	8AO	H	117	0.09	HALF-WAVE RECTIFIER	MAX. A.C. PLATE VOLTS (RMS), 117. MAX. D.C. OUTPUT CURRENT, 75 MA.									
						CLASS A AMPLIFIER	105	−5.2	105	4.0	43.0	17000	5300	——	4000	0.85
117N7-GT	RECTIFIER-BEAM POWER AMPLIFIER	8AV	H	117	0.09	HALF-WAVE RECTIFIER	MAX. A.C. PLATE VOLTS (RMS), 117. MAX. D.C. OUTPUT CURRENT, 75 MA.									
						CLASS A AMPLIFIER	100	−6.0	100	5.0	51.0	16000	7000	——	3000	1.2
117P7-GT	RECTIFIER-BEAM POWER AMPLIFIER	8AV	H	117.0	0.09	HALF-WAVE RECTIFIER	MAX. A.C. PLATE VOLTS (RMS), 117 MAX. D.C. OUTPUT CURRENT, 75 MA.									
						CLASS A AMPLIFIER	105	−5.2	105	5.5	43.0	17000	5300	——	4000	0.85
117Z4-GT	HALF-WAVE RECTIFIER	5AA	H	117.0	0.04	RECTIFIER	MAX. A.C. PLATE VOLTS (RMS), 117 MAX. PEAK INVERSE VOLTS 350				MAX. D.C. OUTPUT MA., 90 MAX. PEAK PLATE MA. 540			MIN. TOTAL EFFECTIVE SUPPLY IMPEDANCE 30 OHMS		
117Z6-GT/G	RECTIFIER DOUBLER	7Q	H	117	0.075	VOLTAGE DOUBLER	MAX. A.C. VOLTS PER PLATE (RMS), 117 MAX. D.C. OUTPUT MA., 60				MIN. TOTAL EFFECTIVE POWER SUPPLY IMPEDANCE: HALF-WAVE, 30 OHMS; FULL-WAVE, 15 OHMS.					
						HALF-WAVE RECTIFIER	MAX. A.C. VOLTS PLATE (RMS), 235 MAX. D.C. OUTPUT MA., PER PLATE, 60				MIN. TOTAL EFFECTIVE POWER SUPPLY IMPEDANCE PER PLATE: UP TO 117 VOLTS, 15 OHMS; AT 150 VOLTS, 40 OHMS; AT 235 VOLTS, 100 OHMS					
1231	TRIPLE-GRID AMPLIFIER	8V	H	6.3	0.45	USED AS PENTODE (5) CLASS A AMPLIFIER	300	200 Ω CATH. RESISTOR	150	2.5	10.0	700000	5500	3850	——	——
						USED AS TETRODE CLASS A AMPLIFIER (30)	300	200 Ω CATH. RESISTOR	150	0.5	12.0	540000	6500	3500	——	——
						USED AS TRIODE CLASS A AMPLIFIER (6)	250	400 Ω CATH. RESISTOR	——	——	13.0	5200	6300	33	——	——
1232	TRIPLE-GRID AMPLIFIER	REFER TO TYPE 7G7/1232 DATA														
1851	TELEVISION AMPLIFIER PENTODE	7R	H	6.3	0.45	CLASS A AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6AC7/1852									
1852	TELEVISION AMPLIFIER PENTODE	REFER TO TYPE 6AC7/1852 DATA														
1853	TELEVISION AMPLIFIER PENTODE	REFER TO TYPE 6AB7/1853 DATA														

MISCELLANEOUS AND SPECIAL-PURPOSE TUBES

TYPE	DESIGN	SOCKET CONN.	CATHODE TYPE AND RATING			USED AS	PLATE SUPPLY VOLTS	GRID BIAS (2) VOLTS	SCREEN SUPPLY VOLTS	SCREEN CURRENT MILLIAMP.	PLATE CURRENT MILLIAMP.	A.C. PLATE RESISTANCE OHMS	TRANSCON-DUCTANCE (GRID-PLATE) µMHOS	AMPLI-FICATION FACTOR	LOAD FOR STATED POWER OUTPUT OHMS	POWER OUTPUT WATTS
			C.T.	VOLTS	AMP.											
2A4-G	ARGON-FILLED THYRATRON	5S	F	2.5	2.5	CONTROL TUBE	200	AVERAGE ANODE CURRENT, 100 MA. PEAK ANODE CURRENT, 1.25 AMP. TUBE DROP, 15 VOLTS. COLD STARTING TIME, 2 SECONDS								
2V3-G	HALF-WAVE RECTIFIER	4Y	F	2.5	5.0	RECTIFIER	MAX. PEAK INVERSE PLATE VOLTS 16500. MAX. PEAK PLATE CURRENT, 12 MA. AVERAGE PLATE CURRENT, 2 MA.									
2X2/879	HALF-WAVE RECTIFIER	4AB	H	2.5	1.75	RECTIFIER	MAX. A.C. PLATE VOLTS (RMS), 4500 MAX. PEAK INVERSE VOLTS, 12500				MAX. D.C. OUTPUT MA., 7.5 MAX. PEAK PLATE MA., 100					
2X3(G)	HALF-WAVE RECTIFIER (41)	4X	F	2.5	2.0	WITH CONDENSER-INPUT FILTER	MAX. A.C. PLATE VOLTS (RMS), 350 MAX. PEAK INVERSE VOLTS 1400				MAX. D.C. OUTPUT MA., 125 MAX. PEAK PLATE MA., 375			MIN. TOTAL EFFECTIVE SUPPLY IMPEDANCE, 10 OHMS		
						WITH CHOKE-INPUT FILTER	MAX. A.C. PLATE VOLTS (RMS), 500 MAX. PEAK INVERSE VOLTS, 1400				MAX D.C. OUTPUT MA. 125 MAX. PEAK PLATE MA., 375			MINIMUM VALUE OF INPUT CHOKE, 5 HENRIES		
2Z2/G84	HALF-WAVE RECTIFIER	4B	F	2.5	1.50	RECTIFIER	MAX. A.C. PLATE VOLTS (RMS), 350				MAX. D.C. OUTPUT CURRENT, 50 MA.					
12A6	BEAM POWER AMPLIFIER	7AC	H	12.6	0.15	CLASS A AMPLIFIER	250	−12.5	250	3.5	30.0	70000	3000	—	7500	2.8
RK-24	TRIODE AMPLIFIER OSCILLATOR	4D	D.C. F	2.0	0.12	A.F. CLASS A AMPLIFIER	90 180	−4.5 −13.5	—	—	4.5 8.0	5500 5000	1450 1600	8 8	5000 12000	0.025 0.250
						R.F. AMPLIFIER OR OSCILLATOR-CLASS C	180	−45.0	—	—	16.5	D.C. GRID CURRENT 6 MA, PEAK R.F. INPUT VOLTAGE, 92. R.F. DRIVING POWER, 0.5 WATT				2.0
RK-42	TRIODE AMPLIFIER OSCILLATOR	4D	D.C. F	1.5	0.06	CLASS A AMPLIFIER	180	−13.5	—	—	3.9	10300	800	8.2	—	—
RK-43	TWIN-TRIODE POWER AMPLIFIER OSCILLATOR	6C	D.C. F	1.5	0.12	CLASS A AMPLIFIER	135	−4.5	—	—	3.0	14500	900	13	VALUES FOR ONE TRIODE	
						CLASS B AMPLIFIER	135	−6.0	—	—	4.0	—	—	—	24000	0.95 (36)
						R.F. AMPLIFIER	135	−20.0	D.C. GRID CURRENT, 3 MA.		14.0	TRIODES CONNECTED IN PUSH-PULL				1.25
M-54	TETRODE POWER AMPLIFIER	TINNED LEADS	D.C. F	0.625	0.04	AMPLIFIER	45	−4.0	45	0.1	0.8	13000	—	—	35000	0.005
RK-62	GAS TRIODE DET. THYRATRON	4D	F	1.4	0.05	SUPER-REGEN.-DETECTOR CONTROL-TUBE	45				1.5	RELAY RESISTANCE, 5000 TO 10000 OHMS. AVERAGE ANODE DROP, 30 VOLTS.				
M-74	TETRODE VOLTAGE AMPLIFIER	TINNED LEADS	D.C. F	0.625	0.02	AMPLIFIER	45	0	22.5	0.07	0.34	500000	125	63	—	—
VR75-30	GAS-FILLED REGULATOR	4W	COLD	—	—	VOLTAGE REGULATOR	MIN. STARTING VOLTAGE, 105 VOLTS. OPERATING VOLTAGE, 75 VOLTS OPERATING CURRENT: MINIMUM, 5 MA.; MAXIMUM, 30 MA.									
G84	HALF-WAVE RECTIFIER	REFER TO TYPE 2Z2/G84 DATA														
VR90-30	GAS-FILLED REGULATOR	4W	COLD	—	—	VOLTAGE REGULATOR	MIN. STARTING VOLTAGE, 125 VOLTS. OPERATING VOLTAGE, 90 VOLTS. OPERATING CURRENT: MINIMUM, 10 MA.; MAXIMUM, 30 MA.									
VR105-30	GAS-FILLED REGULATOR	4W	COLD	—	—	VOLTAGE REGULATOR	MIN. STARTING VOLTAGE, 137 VOLTS. OPERATING VOLTAGE, 105 VOLTS. OPERATING CURRENT: MINIMUM, 5 MA.; MAXIMUM, 30 MA.									

HY-113	MINIATURE TRIODE	SPECIAL	D.C. F	1.4	0.70	OSCILLATOR DETECTOR	45	−4.5	—	—	0.4	25000	250	6.3	—	—
HY-114	TRIODE	SPECIAL	D.F F	1.4	0.12	U.H.F. OSCILLATOR DETECTOR AMPLIFIER	180	OSCILLATOR GRID CURRENT, 3 MA.			15.0	20000	1000	20	—	—
HY-115	MINIATURE PENTODE (31)	SPECIAL	D.F. F	1.4	0.70	VOLTAGE AMPLIFIER	45	−1.5	22.5	0.008	0.03	5 200 000	58	300	—	—
HY-125	MINIATURE PENTODE (31)	SPECIAL	D.C. F	1.4	0.70	A.F. OUTPUT AMPLIFIER	45	−3.0	45	0.2	0.9	—	—	225	50000	0.0115
VR150-30	GAS-FILLED REGULATOR	4W	COLD	—	—	VOLTAGE REGULATOR	MIN. STARTING VOLTAGE, 180 VOLTS. OPERATING VOLTAGE, 150 VOLTS. OPERATING CURRENT; MINIMUM, 5 MA.; MAXIMUM, 30 MA.									
HY-245	PENTODE VOLTAGE AMPLIFIER	HY-245 HY-255	D.C. F	1.25	0.028	CLASS A AMPLIFIER	45	0	45	0.2	0.4	1000000	375	—	—	—
HY-255	PENTODE POWER AMPLIFIER	HY-245 HY-255	D.C. F	1.25	0.028	CLASS A AMPLIFIER	45	−1.5	45	0.35	1.1	—	450	—	—	—
CK-501 (X) (32)	MINIATURE PENTODE (31)	SPECIAL	D.C. F	1.25	0.033	VOLTAGE AMPLIFIER	30 45	0 −1.25	30 45	0.06 0.06	0.3 0.3	1000000 1500000	325 300	—	—	—
CK-502 (X) (32)	MINIATURE PENTODE (31)	SPECIAL	D.C. F	1.25	0.033	A.F. OUTPUT AMPLIFIER (33)	30 45	0 −1.25	30 45	0.06 0.06	0.55 0.6	500 000 700 000	400 500	—	60000 80000	0.0035 0.011
CK-503 (X) (32)	MINIATURE PENTODE (31)	SPECIAL	D.C. F	1.25	0.033	A.F. OUTPUT AMPLIFIER (34)	30	0	30	0.35	1.5	150000	600	—	20000	0.007
CK-504 (X) (32)	MINIATURE PENTODE (31)	SPECIAL	D.C. F	1.25	0.033	A.F. OUTPUT AMPLIFIER (33)	30	0	30	0.09	0.4	500000	350	—	60000	0.0045
CK-505 (X) (32)	MINIATURE PENTODE (31)	SPECIAL	D.C. F	0.625	0.03	IMPEDANCE-COUPLED VOLTAGE AMPLIFIER	30 45	0 −1.25	30 40	0.07 0.08	0.17 0.2	1 100 000 2 000 000	140 150	—	—	—
						RESISTANCE-COUPLED VOLTAGE AMPLIFIER	30	0	30	0.007	0.02	GAIN PER STAGE = 15				
HY-615	TRIODE	SPECIAL	H	6.3	0.15	U.H.F. OSCILLATOR DETECTOR AMPLIFIER	300	OSCILLATOR GRID CURRENT, 3 MA.			20	20000	2200	22	—	4.0
864	NON-MICROPHONIC TRIODE	4D	D.C. F	1.1	0.25	CLASS A AMPLIFIER	90 135	−4.5 −9.0	—	—	2.9 3.5	13500 12700	610 645	8.2 8.2	—	—
874	GAS-FILLED REGULATOR	SPECIAL	COLD	—	—	VOLTAGE REGULATOR	MINIMUM STARTING VOLTAGE, 125 VOLTS. OPERATING VOLTAGE, 90 VOLTS. OPERATING CURRENT: MINIMUM, 10 MA.; MAXIMUM, 50 MA.									
878	HALF-WAVE RECTIFIER	4AB	F	2.5	5.0	RECTIFIER	MAX. PEAK INVERSE PLATE VOLTS, 20000. MAX. A.C. PLATE VOLTAGE (RMS), 7100. MAX. D.C. OUTPUT CURRENT, 5 MA.									
879	HALF-WAVE RECTIFIER	REFER TO TYPE 2X2/879 DATA														
884 885	GAS TRIODE	6Q 5A.	H H	6.3 2.5	0.6 1.4	SWEEP OSCILLATOR	INSTANTANEOUS ANODE VOLTS, 300. MAX. PEAK ANODE CURRENT, 300 MA. AVERAGE ANODE CURRENT: 2-3 MA.									
						GRID-CONTROLLED RECTIFIER	MAX. PEAK VOLTAGE BETWEEN ANY TWO ELECTRODES, 350 VOLTS. MAX. PEAK ANODE CURRENT, 300 MA. MAX. AVERAGE ANODE CURRENT, 75 MA. GRID RESISTOR: LESS THAN 1000 OHMS PER INSTANTANEOUS GRID VOLT.									
954	ACORN PENTODE DETECTOR-AMPLIFIER	1	H	6.3	0.15	CLASS A AMPLIFIER	90 250	−3.0 −3.0	90 100	0.5 0.7	1.2 2.0	1000000 1500000 +	1100 1400	1100 2000 +	—	—
955	ACORN TRIODE DETECTOR-AMPLIFIER-OSCILLATOR	2	H	6.3	0.15	CLASS A AMPLIFIER	90 135 180	−2.5 −3.75 −5.0	—	—	2.5 3.5 4.5	14700 13200 12500	1700 1900 2000	25 25 25	— — 20000	— — 0.135
						CLASS C AMP.—OSC.	180	−35, APPROX.	MAX. GRID CURRENT, 2 MA.		8, MAX.					0.5
956	ACORN SUPER-CONTROL R.F. AMPLIFIER PENTODE	1	H	6.3	0.15	CLASS A AMPLIFIER	250	−3.0	100	1.8	5.5	800000	1800	1440	—	—
						MIXER IN SUPERHETERODYNE	100 250	−10.0 −10.0	100 100	OSCILLATOR PEAK VOLTS = 7.0						

TYPE	DESIGN	SOCKET CONN.	CATHODE TYPE AND RATING			USED AS	PLATE SUPPLY VOLTS	GRID BIAS (2) VOLTS	SCREEN SUPPLY VOLTS	SCREEN CURRENT MILLIAMP.	PLATE CURRENT MILLIAMP.	A.C. PLATE RESISTANCE OHMS	TRANSCON-DUCTANCE (GRID-PLATE) µMHOS	AMPLI-FICATION FACTOR	LOAD FOR STATED POWER OUTPUT OHMS	POWER OUTPUT WATTS
			C.T.	VOLTS	AMP.											
957	ACORN TRIODE DETECTOR–AMPLIFIER–OSCILLATOR	3	D.C. F	1.25	0.05	CLASS A AMPLIFIER	135	−5.0	—	—	2.0	24600	650	16	—	—
958	ACORN TRIODE A.F. AMPLIFIER OSCILLATOR	3	D.C. F	1.25	0.10	CLASS A AMPLIFIER	135	−7.5	—	—	3.0	10000	1200	12	—	—
959	ACORN PENTODE DETECTOR–AMPLIFIER	4	D.C. F	1.25	0.05	CLASS A AMPLIFIER	135	−3.0	67 5	0.4	1.7	800000	600	480	—	—
991	GAS-FILLED REGULATOR	BAYONET CANDELABRA	COLD	—	—	VOLTAGE REGULATOR	MINIMUM STARTING VOLTAGE, 87 VOLTS. OPERATING VOLTAGE, 48 TO 67 VOLTS. OPERATING CURRENT: MINIMUM 0.5 MA.; MAXIMUM, 2 MA.									
1221 1223	NON-MICROPHONIC PENTODE	6F 7R	H	6.3	0.3	CLASS A AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6J7									
1602	NON-MICROPHONIC A.F. AND R.F TRIODE	4D	F	7.5	1.25	CLASS A AUDIO AMPLIFIER	250 350 425	−23.5 −32.0 −40.0	—	—	10 16 18	6000 5150 5000	1330 1550 1600	—	13000 11000 10200	0.4 0.9 1.6
						CLASS B AUDIO AMPLIFIER	250 350 425	−28.0 −40.0 −50.0	—		8.0 (13) 8.0 (13) 8.0 (13)	—	—	—	4000 6000 8000	13.0 (12) 20.0 (12) 25.0 (12)
1603	NON-MICROPHONIC TRIPLE-GRID DETECTOR AMPLIFIER	6F	H	6.3	0.3	DETECTOR–AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6J7									
1611	POWER AMPLIFIER PENTODE	7S	H	6.3	0.7	RELAY CONTROL TUBE	FOR CHARACTERISTICS, REFER TO TYPE 6F6									
1612	NON-MICROPHONIC PENTAGRID MIXER-AMPLIFIER	7T	H	6.3	0.3	MIXER OR CLASS A AMPLIFIER	FOR OTHER CHARACTERISTICS, REFER TO TYPE 6L7									
1620	NON-MICROPHONIC TRIPLE-GRID DETECTOR-AMPLIFIER	7R	H	6.3	0.3	AMPLIFIER	" " " " " " 6J7									
1621	CONTINUOUS-SERVICE POWER AMPLIFIER PENTODE	7S	H	6.3	0.7	USED AS TRIODE (21) PUSH-PULL CLASS A_1 AMPLIFIER	327.5	500 Ω CATH. RESISTOR	—	—	55.0 (13)	—	—	—	5000	2.0 (12)
						USED AS PENTODE PUSH-PULL CLASS A_1 AMPLIFIER	300	−30.0	300	6.5 (13)	38.0 (13)	—	—	—	4000	5.0 (12)
1622	CONTINUOUS-SERVICE BEAM POWER AMPLIFIER	7AC	H	6.3	0 9	PUSH-PULL CLASS A_1 AMPLIFIER	300	−20.0	250	4.0 (13)	86.0 (13)	—	—	—	4000	10.0 (12)
1629	ELECTRON-RAY TUBE	1629	H	12.6	0.15	VISUAL INDICATOR	FOR OTHER CHARACTERISTICS REFER TO TYPE 6E5									
1631	BEAM POWER AMPLIFIER	7AC	H	12.6	0.45	AMPLIFIER	FOR APPLICATIONS CRITICAL AS TO UNIFORMITY OF CHARACTERISTICS. FOR OTHER CHARACTERISTICS REFER TO TYPE 6L6. OPERATING DATA APPLY WITHIN LIMITATION OF MAX. PLATE-DISSIPATION RATING.									
1632	BEAM POWER AMPLIFIER	7AC	H	12.6	0.6	AMPLIFIER	FOR APPLICATIONS CRITICAL AS TO UNIFORMITY OF CHARACTERISTICS. FOR OTHER CHARACTERISTICS REFER TO TYPE 25L6. OPERATING DATA APPLY WITHIN PLATE VOLTAGE AND DISSIPATION LIMITATIONS.									
1633	TWIN-TRIODE AMPLIFIER	8BD	H	25	0.15	AMPLIFIER	FOR APPLICATIONS CRITICAL AS TO MATCHING OF THE TWO TRIODE UNITS. FOR OTHER CHARACTERISTICS REFER TO TYPE 12SN7-GT.									

1634	TWIN-TRIODE AMPLIFIER	8S	H	12.6	0.15	AMPLIFIER	FOR APPLICATIONS CRITICAL AS TO MATCHING OF THE TWO TRIODE UNITS. FOR OTHER CHARACTERISTICS REFER TO TYPE 12SC7.									
2050	GAS TETRODE	2050-2051	H	6.3	0.6	GRID-CONTROLLED RECTIFIER	MAX. PEAK FORWARD ANODE VOLTAGE, 650 VOLTS. MAX. PEAK ANODE CURRENT, 500 MA. MAXIMUM AVERAGE ANODE CURRENT, 100 MA. SHIELD GRID (#2) VOLTAGE, 0 VOLTS. MAX. PEAK INVERSE ANODE VOLTAGE, 1300 VOLTS									
2051	GAS TETRODE	2050-2051	H	6.3	0.6	GRID-CONTROLLED RECTIFIER	MAX. PEAK FORWARD ANODE VOLTAGE, 350 VOLTS. MAX. PEAK ANODE CURRENT, 375 MA. MAXIMUM AVERAGE ANODE CURRENT, 75 MA. SHIELD GRID (#2) VOLTAGE, 0 VOLTS. MAX. PEAK INVERSE ANODE VOLTAGE, 700 VOLTS									
9001	DETECTOR AMPLIFIER PENTODE	9001-9003	H	6.3	0.15	CLASS A AMPLIFIER	90 250	−3.0 −3.0	90 100	0.5 0.7	1.2 2.0	1000000 1000000+	1100 1400	—	—	—
						MIXER IN SUPERHETERODYNE	100 250	−5.0 −5.0	100 100	—	—	—	—	CONVERSION TRANSCONDUCTANCE 550 MICROMHOS		
9002	DETECTOR AMPLIFIER TRIODE	9002	H	6.3	0.15	CLASS A AMPLIFIER	90 180 250	−2.5 −5.0 −7.0	—	—	2.5 4.5 6.3	14700 12500 11400	1700 2000 2200	25 25 25	—	—
9003	SUPER-CONTROL AMPLIFIER PENTODE	9001-9003	H	6.3	0.15	CLASS A AMPLIFIER	250	−3.0	100	2.7	6.7	700000	1800	—	—	—
						MIXER IN SUPERHETERODYNE	100 250	−10.0 −10.0	100 100	—	—	—	—	CONVERSION TRANSCONDUCTANCE 600 MICROMHOS		

FOOTNOTE REFERENCES FOR STANDARD AND SPECIAL RECEIVING TUBES

[1] For grid leak detection, plate volts 45, grid return to plus filament.

[2] Either a.c. or d.c. may be used on the filament or heater, except as specifically noted. For use of d.c. on filament types, decrease stated grid volts by ½ of filament voltage.

[3] Supply voltage applied through 20,000-ohm dropping resistor.

[4] Mercury vapor type.

[5] Grid no. 1 is control grid; grid no. 2 is screen; grid no. 3 is tied to cathode.

[6] Grid no. 1 is control grid. Grids nos. 2 and 3 tied to plate.

[7] Grids nos. 1 and 2 connected together; grid no. 3 connected to plate.

[8] Grids nos. 3 and 5 are screen. Grid no. 4 is control grid (input).

[9] Grids nos. 2 and 4 are screen. Grid no. 1 is control grid (input).

[10] For grid of following tube.

[11] Both grids connected together; likewise both plates.

[12] Power output is for 2 tubes at stated plate-to-plate load.

[13] For 2 tubes.

[14] Preferably obtained by using 70,000-ohm dropping resistor in series with 90-volt supply.

[15] Grids nos. 2 and 3 tied to plate.

[16] Applied through plate resistor of 250,000 ohms or 500-hy. choke shunted by 250,000-ohm resistor.

[17] Applied through plate resistor of 100,000 ohms.

[18] Applied through plate resistor of 250,000 ohms.

[19] 50,000 ohms.

[20] Requires different socket from small 7 pin.

[21] Grid no. 3 tied to plate.

[22] Plate voltages greater than 125 volts r.m.s. require 100-ohm (min.) series plate resistor.

[23] Applied through plate resistor of 150,000 ohms.

[24] For signal input control grid. Grid no. 3 bias, minus 3 volts.

[25] Applied through 200,000-ohm plate resistor.

[26] Grids nos. 2 and 4 are screen. Grid no. 3 is control grid.

[27] Maximum.

[28] Megohms.

[29] Grids nos. 1 and 2 tied together.

[30] Grids nos. 2 and 3 tied together.

[31] Designed especially for hearing aid use.

[32] "X" types have removable octal base.

[33] Operates into crystal earphone.

[34] Operates into magnetic reproducer.

[35] Unless otherwise specified, values are for the two units.

[36] Power output is for one tube at stated plate-to-plate load.

[37] Per plate.

[38] Two sections have common plate; value is for each triode.

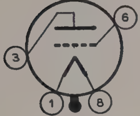
[39] Values are for each unit.

[40] D.c. resistance in grid circuit should not exceed 1.0 megohm under maximum rated conditions per unit.

[41] Values are for two tubes with filaments in series; equivalent to one type 5Y3-GT/5Y3-G.

Subscript 1 on class of amplifier service indicates that grid current does not flow on any part of input cycle.

Subscript 2 on class of amplifier service indicates that grid current flows on some part of input cycle.



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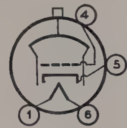

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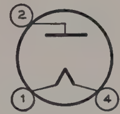

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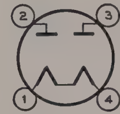

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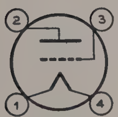

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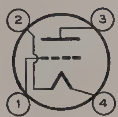

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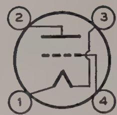

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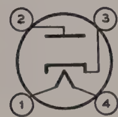

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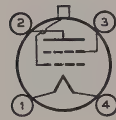

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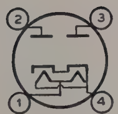

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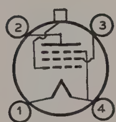

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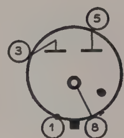

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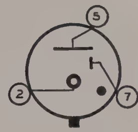

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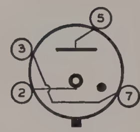

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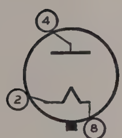

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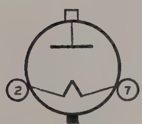

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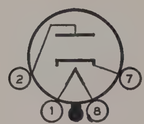

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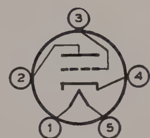

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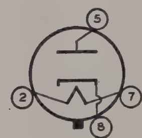

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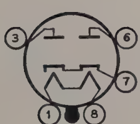

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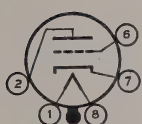

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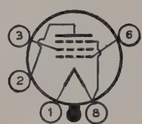

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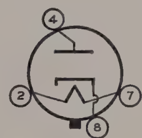

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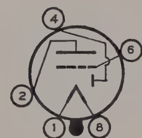

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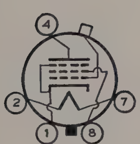

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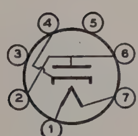

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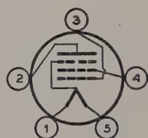

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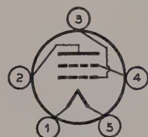

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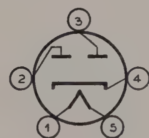

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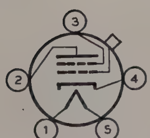

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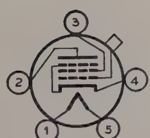

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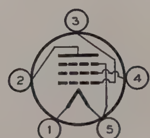

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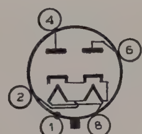

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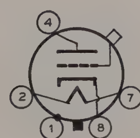

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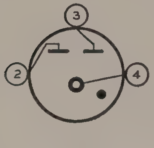
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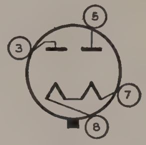
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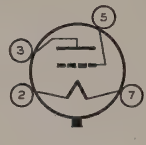
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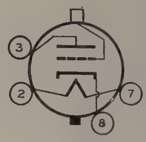
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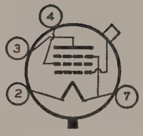
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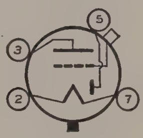
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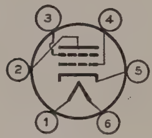
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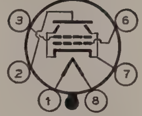
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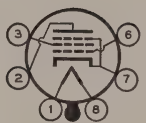
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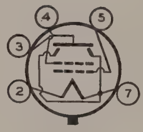
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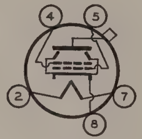
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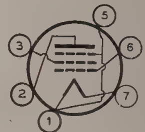
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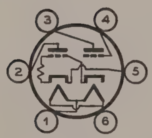
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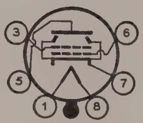
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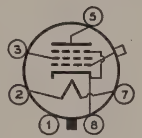
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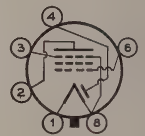
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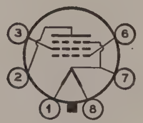
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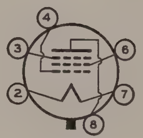
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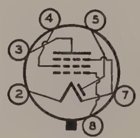
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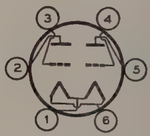
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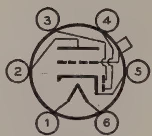
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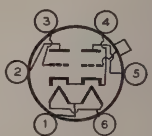
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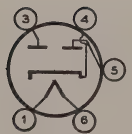
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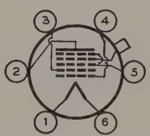
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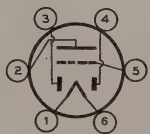
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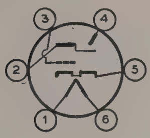
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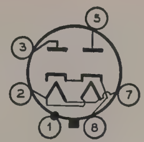
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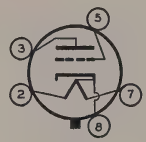
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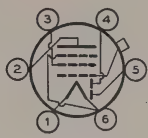
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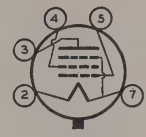
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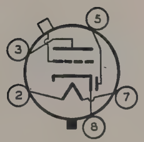
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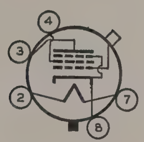
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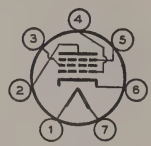
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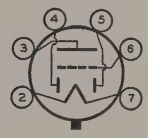
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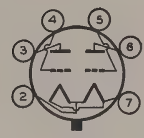
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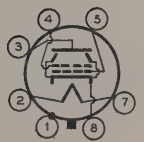
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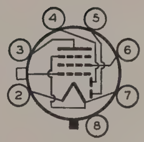
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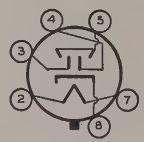
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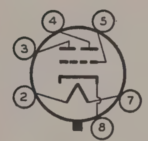
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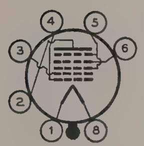
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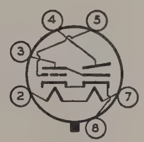
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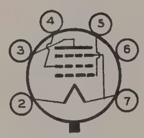
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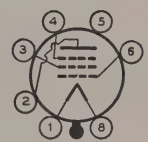
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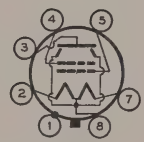
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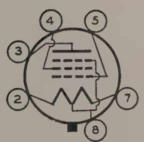
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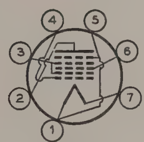
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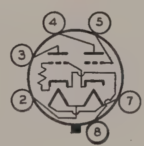
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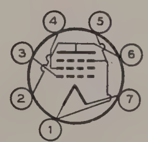
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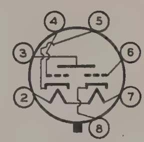
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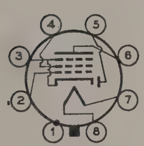
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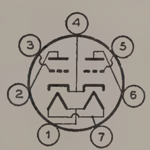
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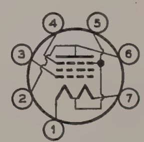
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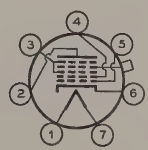
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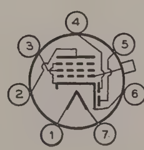
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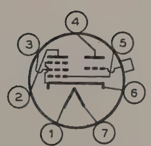
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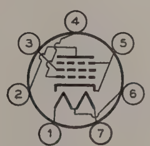
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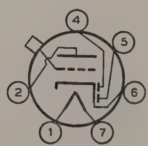
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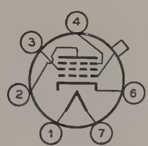
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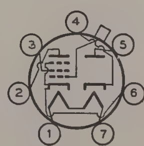
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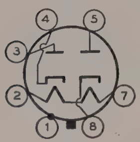
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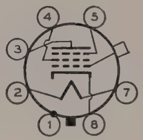
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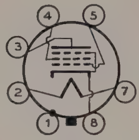
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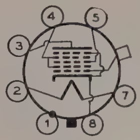
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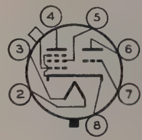
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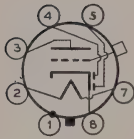
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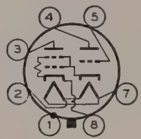
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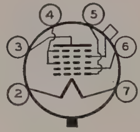
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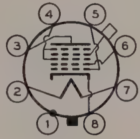
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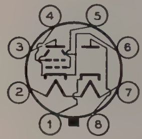
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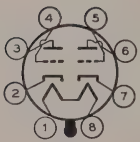
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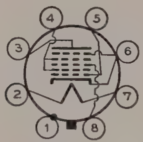
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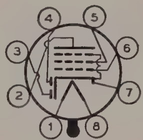
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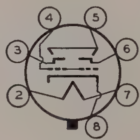
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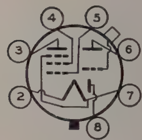
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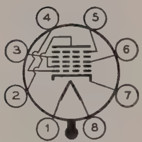
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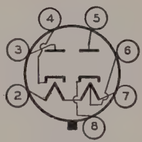
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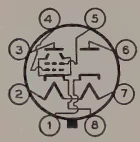
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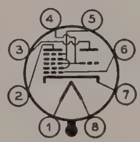
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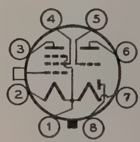
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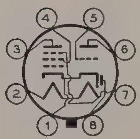
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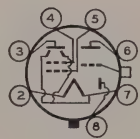
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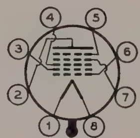
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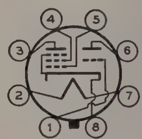
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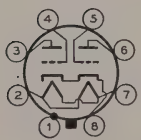
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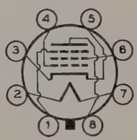
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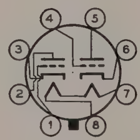
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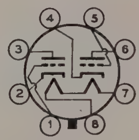
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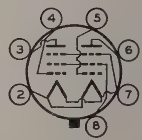
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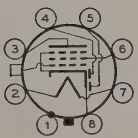
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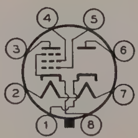
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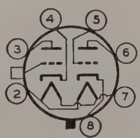
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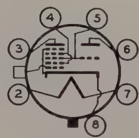
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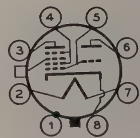
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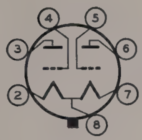
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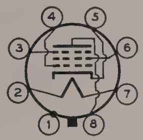
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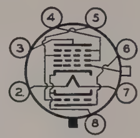
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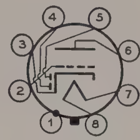
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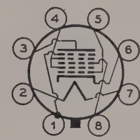
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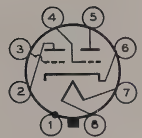
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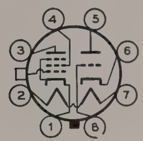
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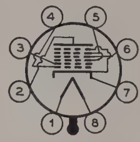
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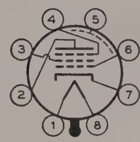
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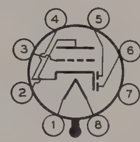
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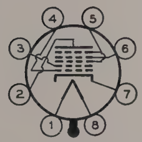
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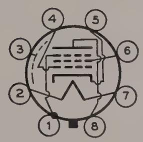
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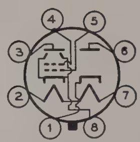
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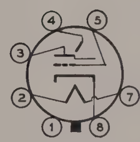
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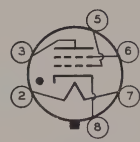
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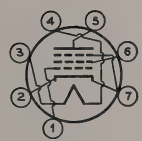
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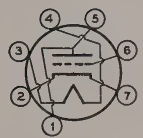
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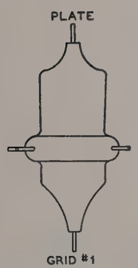

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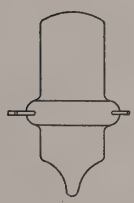
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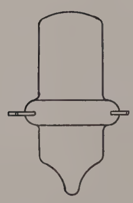
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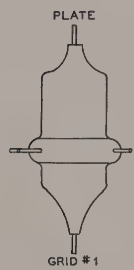

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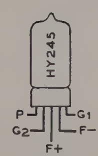

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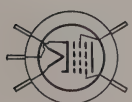
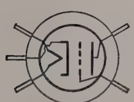

GRID-END VIEW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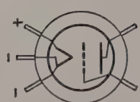
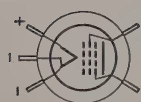

SHORT-END VIEW


GRID-END VIEW

CATHODE-RAY TUBES

OSCILLOSCOPE AND TELEVISION RECEIVING TYPES

TYPE	SOCKET CONN.	HEATER VOLTS	HEATER AMP.	DEFLEC-TION	SCREEN DIA. INCHES	USED AS	ANODE Nº 1 VOLTS	ANODE Nº 2 VOLTS	GRID Nº 1 CUTOFF VOLTS	GRID Nº 2 VOLTS	MAXIMUM PEAK VOLTS BETWEEN ANODE Nº2 AND ANY DEF PLATE	MAXIMUM SCREEN INPUT POWER PER SQUARE CENTIMETER MILLIWATTS	DEFLECTION SENSITIVITY MM./VOLT D.C. D_1 & D_2	DEFLECTION SENSITIVITY MM./VOLT D.C. D_3 & D_4	GRID SIGNAL-SWING VOLTS	SCREEN ① MATERIAL
3AP1/906-P1	1	2.5	2.1	ELECTRO-STATIC	3	OSCILLOSCOPE	128 170 230 285 345 475	400 600 800 1000 1200 1500	APPROX. 20% OF ANODE Nº1 VOLTAGE	—	600	10	0.81 0.55 0.41 0.33 0.27 0.22	0.87 0.58 0.44 0.35 0.29 0.23	—	P1
3AP4/906-P4	1	2.5	2.1	ELECTRO-STATIC	3	PICTURE TUBE	FOR OTHER CHARACTERISTICS, REFER TO TYPE 3AP1/906 P1									P4
5AP1/1805-P1	2	6.3	0.6	ELECTRO-STATIC	5	PICTURE TUBE	430 575	1500 2000	−65 APPROX. −86 APPROX.	—	500	10	0.23 0.17	0.28 0.21	15 20	P1
5AP4/1805-P4	2	6.3	0.6	ELECTRO-STATIC	5	PICTURE TUBE	FOR OTHER CHARACTERISTICS REFER TO TYPE 5AP1/1805-P1									P4
5BP1/1802-P1	2	6.3	0.6	ELECTRO-STATIC	5	OSCILLOSCOPE	250 310 425	1200 1500 2000	APPROX. 20% OF ANODE Nº1 VOLTAGE	—	500	10	0.50 0.40 0.30	0.55 0.44 0.33	—	P1
5BP4/1802-P4	2	6.3	0.6	ELECTRO-STATIC	5	PICTURE TUBE	FOR OTHER CHARACTERISTICS, REFER TO TYPE 5AP1/1802-P1									P4
7AP4	3	2.5	2.1	ELECTRO-MAGNETIC	7	PICTURE TUBE	675	3500	−67.5 APPROX.	—	—	2.5	—	—	15	P4
9AP4/1804-P4	4	2.5	2.1	ELECTRO-MAGNETIC	9	PICTURE TUBE	1225 1425	6000 7000	−75 APPROX. −75 APPROX.	250 250	—	10	—	—	25 25	P4
9CP4	4AF	2.5	2.1	ELECTRO-MAGNETIC	9	PICTURE TUBE	—	7000	110	—	—	10	—	—	25	P4
12AP4/1803-P4	4	2.5	2.1	ELECTRO-MAGNETIC	12	PICTURE TUBE	1240 1460	6000 7000	−75 APPROX. −75 APPROX.	250 250	—	10	—	—	25 25	P4
12CP4	4AF	2.5	2.1	ELECTRO-MAGNETIC	12	PICTURE TUBE	—	7000	110	—	—	10	—	—	25	P4
902	5	6.3	0.6	ELECTRO-STATIC	2	OSCILLOSCOPE	100 150	400 600	−80 APPROX.	—	350	5	0.28 0.19	0.33 0.22	—	P4
903	4	2.5	2.1	ELECTRO-MAGNETIC	9	OSCILLOSCOPE	195 580 900 1360	1000 3000 4600 7000	— — — −120 APPROX.	100 250 250 250	—	10	—	—	—	P1
904	6	2.5	2.1	ELECTRO-STATIC—MAGNETIC	5	OSCILLOSCOPE	210 630 970	1000 3000 4600	— — −140 APPROX.	100 100 250	4000	10	0.40 0.13 0.09	— —	—	P1
905	7	2.5	2.1	ELECTRO-STATIC	5	OSCILLOSCOPE	225 450	1000 2000	— −60 APPROX.	—	1000	10	0.38 0.19	0.46 0.23	—	P1
907	7	2.5	2.1	ELECTRO-STATIC	5	OSCILLOSCOPE	FOR OTHER CHARACTERISTICS, REFER TO TYPE 905									P5
908	1	2.5	2.1	ELECTRO-STATIC	3	OSCILLOSCOPE	FOR OTHER CHARACTERISTICS, REFER TO TYPE 3AP1/906-P1									P5

909	7	2.5	2.1	ELECTRO-STATIC	5	OSCILLOSCOPE	FOR CHARACTERISTICS, REFER TO TYPE 905									P2
910	1	2.5	2.1	ELECTRO-STATIC	3	OSCILLOSCOPE	FOR CHARACTERISTICS, REFER TO TYPE 3AP1/906-P1									P2
911	1	2.5	2.1	ELECTRO-STATIC	3	OSCILLOSCOPE	FOR CHARACTERISTICS, REFER TO TYPE 3AP1/906-P1 [2]									
912	8	2.5	2.1	ELECTRO-STATIC	5	OSCILLOSCOPE	1000 2000 3000	5000 10000 15000	— — −125 APPROX.	250 250 250	7000	10	0.083 0.041 0.028	0.102 0.051 0.034	—	P1
913	5	6.3	0.6	ELECTRO-STATIC	1	OSCILLOSCOPE	50 100	250 500	— −90	—	250	5	0.15 0.07	0.21 0.10	—	P1
914	9	2.5	2.1	ELECTRO-STATIC	9	OSCILLOSCOPE	460 915 1500	2500 5000 7000	−100 APPROX. −100 APPROX. −100 APPROX.	250 250 250	3000	10	0.204 0.102 0.073	0.260 0.130 0.093	—	P1
1800	4	2.5	2.1	ELECTRO-MAGNETIC	9	PICTURE TUBE	625 925 1250	3000 4500 6000	— — −75 APPROX.	200 250 250	—	10	—	—	20 25 25	P3
1801	10	2.5	2.1	ELECTRO-MAGNETIC	5	PICTURE TUBE	375 450	2500 3000	— −35 APPROX.	—	—	10	—	—	15 20	P3

CATHODE-RAY TRANSMITTING TYPES

TYPE	USED AS	SOCKET CONN.	HEATER VOLTS	HEATER AMP.	ANODE No 1	ANODE No 2	GRID No 1 CUT-OFF VOLTS	GRID No 2 VOLTS	COLLECTOR VOLTS	COLLECTOR µAMPS.	AVERAGE D.C. DEF. PLATE VOLTS	ROTATOR ELECTRODE VOLTS	MASK ELECTRODE VOLTS	DEFLECTING FLUX DENSITY GAUSSES	FOCUSING DENSITY GAUSSES	PEAK-TO-PEAK DEFLECTING VOLTAGE HORIZONTAL	PEAK-TO-PEAK DEFLECTING VOLTAGE VERTICAL	TYPE OF PICKUP
1840	ORTHICON	11	6.3	0.6	250	—	−40 APPROX.	225	—	—	225	100 APPROX.	−3	25 APPROX.	70 APPROX.	160	—	DIRECT OR FILM
1847	ICONOSCOPE	12	6.3	0.6	150	600 [3]	−120 APPROX.	600 [3]	600 [3]	—	—	—	—	—	—	200	225	DIRECT
1848	ICONOSCOPE	13	6.3	0.6	300 APPROX.	1000 [4]	−50 APPROX	1000	1000 [4]	0.1	—	—	—	—	—	—	—	DIRECT
1849	ICONOSCOPE	14	6.3	0.6	360	1000	−30 APPROX.	—	1000	0.05 TO 0.1	—	—	—	—	—	—	—	FILM
1850	ICONOSCOPE	14	6.3	0.6	FOR OTHER CHARACTERISTICS, REFER TO TYPE 1849													DIRECT
1898	MONOSCOPE	15	2.5	2.1	240 300 360	800 1000 1200	−50 APPROX. −60 APPROX. −70 APPROX.	PATTERN ELECTRODE VOLTAGE 750 950 1150					BEAM CURRENT (µAMP.) 1 APPROX. 2 APPROX. 3 APPROX.			135 170 200	125 155 185	TEST PATTERN
1899	MONOSCOPE	16	2.5	2.1	260 390	1000 1500	— −60	—	1050 1700	— —	PATTERN ELECTRODE VOLTAGE 1000 1500				BEAM CURRENT (µAMP.) 2 4			TEST PATTERN

REFERENCES

[1] Screen materials are classified as follows: Phosphor no. 1 is of medium persistence and produces green fluorescence. Phosphor no. 2 is of long persistence and produces bluish-white fluorescence. Phosphor no. 3 is of medium persistence and produces yellow fluorescence. Phosphor no. 4 is of medium persistence and produces white fluorescence. Phosphor no. 5 is of short persistence and produces bluish fluorescence.

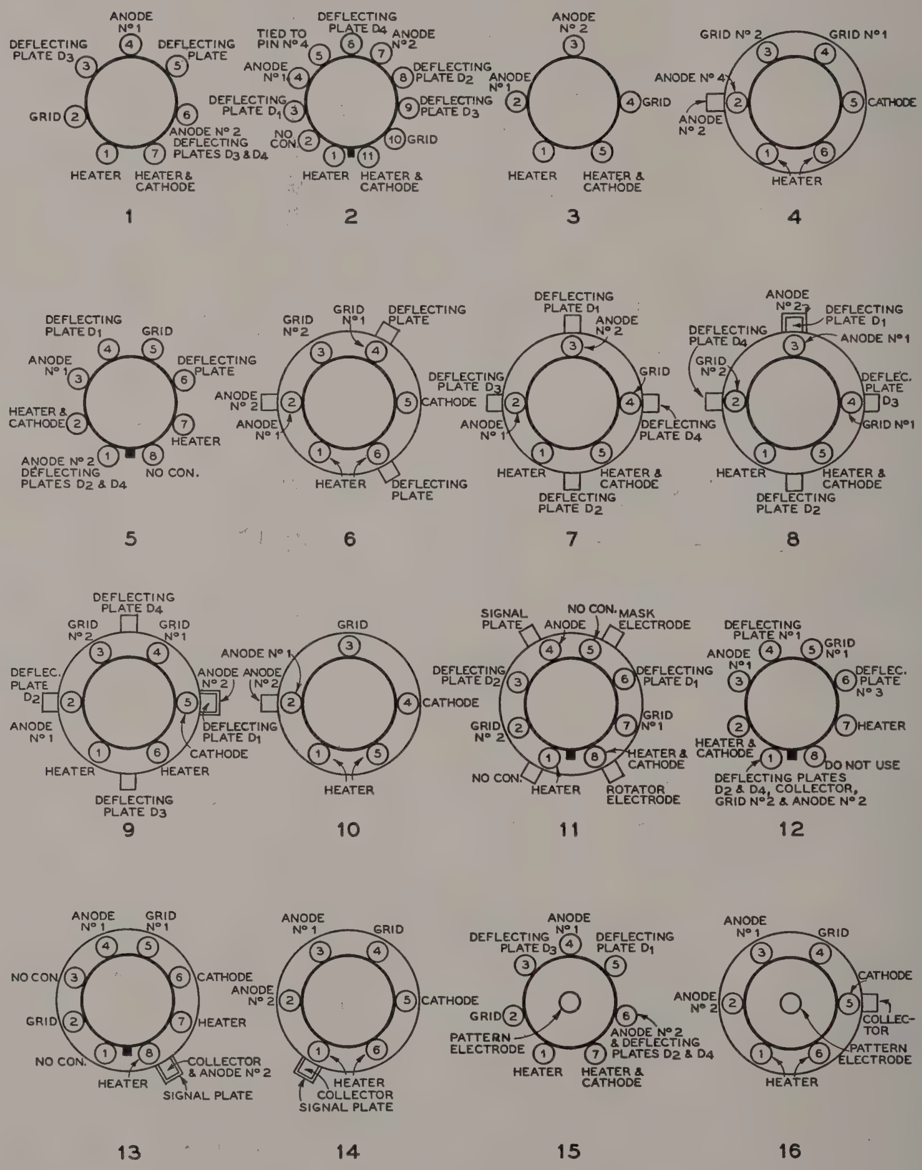
[2] The electron gun of the 907 is designed to be unusually free from magnetization effects.

[3] Collector, grid no. 2, and anode no. 2 are connected together within the tube.

[4] Collector and anode no. 2 are connected together within the tube.

CATHODE RAY TUBE SOCKET CONNECTIONS

BOTTOM VIEWS



CHAPTER SIX

Radio Receiver Construction

THE receivers to be described in this chapter can, for the most part, be constructed with a few inexpensive hand tools. Whether one saves anything over purchasing a factory built receiver depends upon several factors (see Chapter 25). In any event, there is the satisfaction of constructing one's own equipment, and the practical experience that can be gained only by actually building apparatus.

After finishing the wiring of these receivers it is suggested that one go over the wiring very carefully to check for errors before applying plate voltage to the receiver. If possible, have someone else check the wiring after you have gone over it yourself. Some tubes can be damaged permanently by having screen voltage applied when there is no voltage on the plate. Electrolytic condensers can be damaged permanently by hooking them up backwards (wrong polarity). Transformer, choke, and coil windings can be burned out by incorrect wiring of the high voltage leads. Almost any tube can be damaged by hooking up the elements incorrectly; no tube can last long with plate voltage applied to the control grid.

Before starting construction, it is suggested that one read the chapter on *Workshop Practice* (Chapter 25).

SIMPLE 2-TUBE AUTODYNE

A simple yet versatile receiver of modest cost is illustrated in Figures 1, 2, and 4. The receiver uses an autodyne detector and one stage of impedance coupled a.f. to give good earphone volume on all signals. The circuit is quite simple, as inspection of Figure 3 will disclose.

The receiver uses 6.3-volt tubes, which may be supplied heater power from either a small 6.3-volt filament transformer or a regular 6-volt auto battery. For regular home use a transformer is recommended, but the provision for use with a battery permits semi-portable operation. This makes the receiver a good one for a beginner, as it can be used

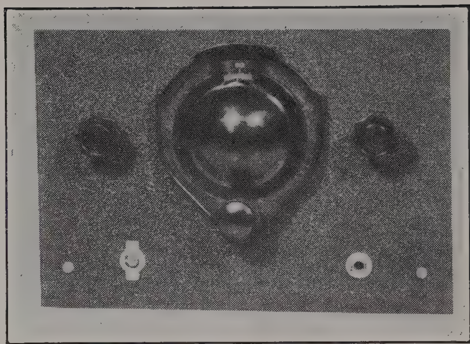


Figure 1.
SIMPLE 2-TUBE AUTODYNE RECEIVER.
This receiver is inexpensive to build and has excellent weak signal response. While not as selective as more elaborate receivers, it makes a good set for the newcomer's first receiver.

as a portable or emergency receiver later on, should one decide to build or buy a more elaborate receiver.

Plate voltage is supplied from a standard medium-duty 45-volt B battery. Such a battery, costing only a little over a dollar, will last over a year with normal use, as the B current drain of the receiver is only a few milliamperes. This voltage is sufficient for good performance of the receiver, because the full plate voltage is supplied to the detector as a result of the use of a choke (CH_1) instead of the usual plate resistor in the plate circuit of the detector. Also, the *amplification* of the 6C5 is practically as great at 45 volts as at the full maximum rated voltage of 250 volts. The maximum undistorted power output of the a.f. stage is considerably less at 45 volts, but as it is more than sufficient to drive a pair of phones, there is no point in using higher plate voltage. For these reasons, a single B battery was decided upon in preference to an a.c. power pack, because the battery

is not only much less expensive but also permits portable operation.

When wired as shown in the diagram, the receiver should not be used with higher plate voltage, because the screen potentiometer is across the full plate voltage, and also because the 1¼-volt bias on the 6C5 is not sufficient for higher plate voltage.

The receiver can be built for about $12, including B battery and midget filament transformer, provided inexpensive components are chosen.

While the receiver will operate on 10 meters and a 10-meter coil is included in the coil table, the receiver is designed primarily for 20-, 40-, and 80-meter operation. No matter how well constructed, an autodyne receiver is not particularly effective on 10 meters, especially for 'phone reception. No provision is made for 160-meter operation, as the receiver does not have sufficient selectivity to distinguish between several very loud 'phone signals in the same part of the band.

For 20-, 40-, and 80-meter operation the receiver compares favorably with the most expensive when it comes to picking up weak, distant stations, especially on c.w. However, in common with all autodyne receivers, loud local signals have a tendency to block, and therefore more trouble will be experienced with QRM than with a superheterodyne.

The chassis consists of a 6 x 9-inch Masonite "Presdwood" top and a 1¾-inch back of the same material. These are fastened to two pieces of wood which form the sides of the

COIL TABLE
For 2-Tube Autodyne

All coils wound with no. 22 d.c.c. on standard 1½-inch forms

80 Meters
29 turns close-wound; cathode tap 1½ turns from ground

40 Meters
16 turns spaced 1¾ inches; cathode tap 1½ turns from ground

20 Meters
7 turns spaced 1¼ inches; cathode tap 1½ turns from ground

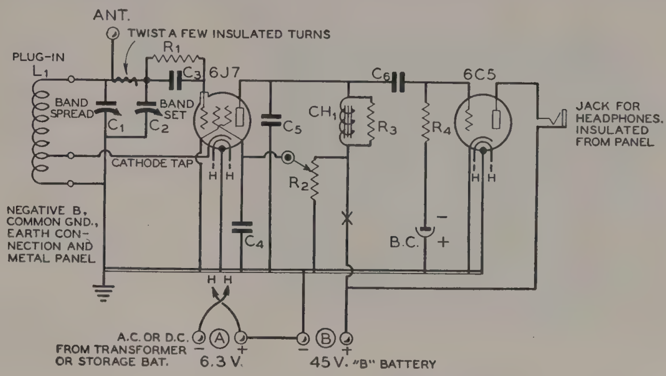
10 Meters
4 turns spaced 1¼ inches; cathode tap 1 turn from ground

chassis. The wooden sides are 1¾ inches high, ¾ inch thick, and are 6 inches long, *including* the Masonite back. The whole thing is held together with wood screws, as may be seen in Figures 2 and 4. A 7 x 11-inch metal front panel is attached to the chassis by means of wood screws sunk in the wooden end pieces of the chassis.

Inexpensive wafer sockets are used. Because the thickness of the chassis would make it necessary to drill holes large enough to take the whole tube base if the sockets were mounted below the chassis, as is customary

Figure 2.
BACK VIEW OF THE 2-TUBE AUTODYNE.
The chassis is made of wood and Masonite wall board. The "shield hat" for the grid leak and condenser hides most of the main tuning condenser.

C_1 — 15-$\mu\mu$fd. midget variable
C_2—100-$\mu\mu$fd. midget variable
C_3—.0001-μfd. smallest size mica condenser
C_4—0.25-μfd. tubular, 400 volts
C_5—.0005-μfd. midget mica
C_6 — .01-μfd. tubular, 400 volts
R_1—3 meg., ½ watt
R_2 — 50,000-ohm potentiometer
R_3—250,000 ohms, ½ watt
R_4—500,000 ohms, ½ watt



CH—300 or more hy., 5 ma.
BC—1¼-volt bias cell
L_1—See coil table

Figure 3.
WIRING DIAGRAM OF 2-TUBE AUTODYNE.
By substituting a 6S7 for the 6J7 and a 6L5-G for the 6C5, the receiver can be run economically from dry cells for heater power. Only 4½ volts is required, and three no. 6 dry cells will give over 150 hours life.

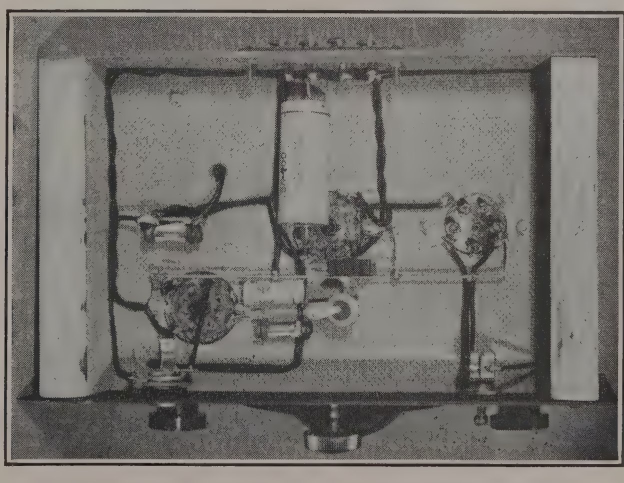
with metal chassis, the sockets are mounted on *top* of the chassis. This is clearly illustrated in the photographs.

Correct connection of the socket terminals can be assured by referring to the socket connections for the 6J7 and 6C5 in Chapter 5. Bear in mind that these are bottom views of the sockets, with the socket facing you the same as when soldering to the terminals from the underside of the chassis.

Connections for filament and plate power are made by means of a terminal strip which is mounted over a hole cut in the back of the chassis. If the proper tools for cutting out a long, rectangular hole are not available, four separate holes about ⅜ inch in diameter will take the terminal screws and lugs. If desired, the terminal strip can be replaced by four Fahnestock clips screwed directly to the back of the chassis.

The phone jack is shown mounted on the front panel, along with a toggle switch in the B plus lead. If mounted on the metal front panel, the phone jack must be insulated from the panel by means of fiber washers to prevent shorting the plate voltage. The jack can be mounted on the back of the chassis, in which case it will not require insulating washers.

Figure 4.
UNDER-CHASSIS VIEW OF 2-TUBE AUTODYNE.
The construction of the chassis and placement of components is clearly illustrated. If desired the phone jack may be mounted on the back of the chassis.



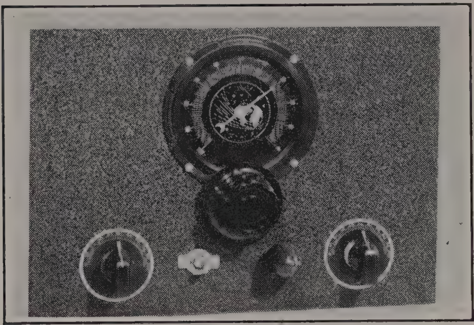


Figure 5.
SIMPLE 3-TUBE SUPERHETERODYNE.
The bandset condenser is to the left, the detector "resonating" condenser to the right. The latter makes an effective volume control. The small knob operates the regeneration potentiometer.

The screen potentiometer is across the B battery and draws a small amount of current even with the filaments turned off; hence, it is necessary either to unhook the B battery when the set is not in use, or else incorporate a switch to accomplish the same thing. If desired, a potentiometer with an "off switch" can be used, in which case the B battery may be disconnected simply by turning the potentiometer knob all the way to the left. The heaters are turned off by turning off the 110-volt supply to the filament transformer.

As is true with any grid leak type detector, the grid lead (including the grid leak and condenser) must be shielded thoroughly in order to avoid bad hum pickup, commonly known as "grid hum." This is accomplished effectively by soldering the grid leak and grid condenser (both of the smallest physical size procurable) directly to the grid clip, and shielding the whole business by means of a "hat" consisting of a regular metal tube grid shield cap to which is soldered a rectangular piece of tin can or galvanized iron as shown in the illustration. The latter measures about $1\frac{1}{2}$ x 3 inches and is bent in the form of a "U," then soldered to the grid clip shield. Care must be taken that the shield does not short out against any of the connecting leads.

The antenna may consist of a 50 to 100 foot length of wire as high and in the clear as possible. It is capacity coupled to the receiver by means of a few turns of insulated wire around the grid lead. A small 3-30 $\mu\mu$fd. compression type mica trimmer may be substituted for the twisted wire as a variable coupling condenser, if desired.

After the correct position of the bandset condenser (C_2) is determined for a given band, a scratch or mark is made on the back rotor plate to enable one to adjust the bandset condenser for any band simply by observing the marks on the bandset condenser.

The wiring diagram assumes that the receiver will be used with magnetic type earphones. If crystal earphones are used, a small 30-hy. choke should be connected across the headphone jack.

SIMPLE 3-TUBE SUPERHETERODYNE

The small superheterodyne shown in the accompanying illustrations has many of the advantages of sets having many more tubes. It has good image rejection, selectivity and sensitivity, and drives either phones or a dynamic loudspeaker to good volume.

A 6K8 converter directly feeds a regenerative second detector operating at a frequency just above 1500 kc. The latter is impedance coupled to a beam tetrode audio tube. The plate current and audio power output are too great for a pair of phones; so the phones are connected in the screen circuit.

Excellent selectivity and sensitivity are obtained on 'phone by running up the regeneration on the second detector right to the edge of oscillation. By advancing the regeneration control still farther, the second detector will oscillate, thus providing autodyne reception of code signals. The regeneration also acts as a sensitivity control to prevent blocking by very loud local signals. To keep loud 'phone signals from blocking, the regeneration is decreased way below the edge of oscillation. To keep loud c.w. signals from blocking, the regeneration control is advanced full on.

The 6K8 converter is conventional, and no special precautions need be taken with this stage except to keep the mixer-section leads as short as possible in order to obtain maximum performance on 10 meters. A minimum number of coils is required for all-band operation (10 to 160 meters) because the oscillator coil for each band serves as the detector coil for the next higher frequency band, the tickler serving as the antenna winding. Thus all coils except the 160-meter mixer and 10-meter oscillator coils do double duty.

The set is built on a metal chassis measuring $2\frac{1}{2}$ x 6 x 8 inches. This supports a 7 x 10-inch front panel. The correct placement of components may be determined by referring to the illustrations.

To obtain regeneration in the grid-leak type second detector, a tickler coil is added to the

i.f. transformer. Inspection of Figure 8 will show that the second detector then resembles the common "autodyne" grid-leak detector with regeneration control.

For maximum performance, the detector should go into oscillation when the screen voltage is about 35 volts. This is accomplished by using as a tickler 3 turns of no. 22 d.c.c. wire wound around the dowel of the i.f. transformer, right against the grid winding. Few tickler turns are required, as there is no antenna to load the detector, and therefore it goes into oscillation with but little feedback.

To wind the tickler, simply remove the shield from the i.f. transformer, and, using a 1-foot piece of the same d.c.c. wire used to wind the plug-in coils, wrap 3 turns around the dowel as closely as possible to the grid winding. Then twist the two leads together to keep the turns in place and replace the shield. The polarity of the tickler must be correct for regeneration; if oscillation is not obtained, reverse the two tickler leads.

Care must be taken with the grid leak, grid condenser, and grid lead of the 6SJ7; otherwise there will be "grid hum." The outside foil of the tubular grid condenser should go to the i.f. grid coil and *not* to the grid of the tube. The connection to the grid pin of the 6SJ7 socket should be kept as short as possible—not over $\frac{1}{2}$ inch, and both grid leak and grid condenser should be kept at least $\frac{1}{2}$ inch from other wiring. In some cases it may be necessary to shield the grid leak and condenser with a small piece of grounded tin in order to eliminate grid hum completely.

The phone jack is a special type, commonly called a 2-circuit "filament lighting" jack. It is connected so that when the phones are inserted they not only are connected in the screen circuit in such a way that no d.c. flows through the phones, but the speaker transformer is shorted out in order to silence the speaker. Switching the plate of the 6V6 directly to B plus also improves the quality in the phones slightly.

Any well-filtered power supply delivering between 300 and 375 volts at 50 ma. can be used to supply the receiver. If the speaker is of the p.m. type, requiring no field supply, a 200 to 250 volt power pack will suffice.

Either a 2-wire feeder or single-wire antenna worked against ground can be used. For doublet input, connect to the two antenna coil terminals. For Marconi input, ground one terminal and connect the antenna to the other.

Adjusting the mica trimmer on the grid side of the i.f. transformer changes the intermediate frequency. The trimmer on the plate coil should always be resonated for maximum signal strength. It need not be touched after the initial adjustment unless the grid trimmer is changed. The intermediate frequency should be adjusted to about 1550 kc. and then a check made to make sure it is not right on some nearby broadcast station.

Figure 6.
REAR VIEW OF THE SIMPLE SUPER.
The detector coil is to the left, directly above the detector tuning condenser, and the oscillator coil is to the right. Antenna terminals, power socket, speaker plug socket, and earphone jack may be seen on the backdrop of the chassis.

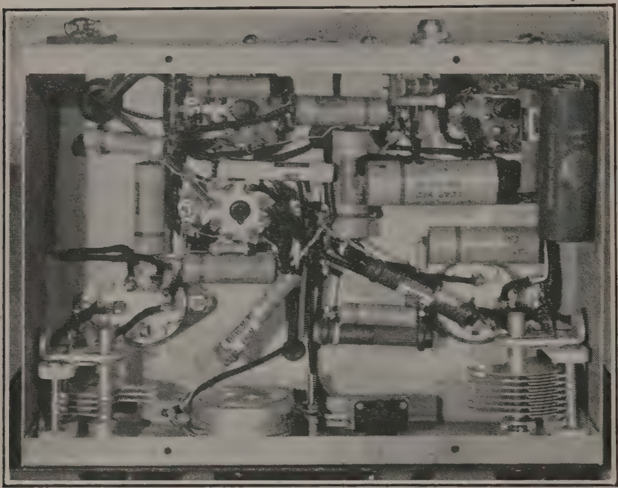
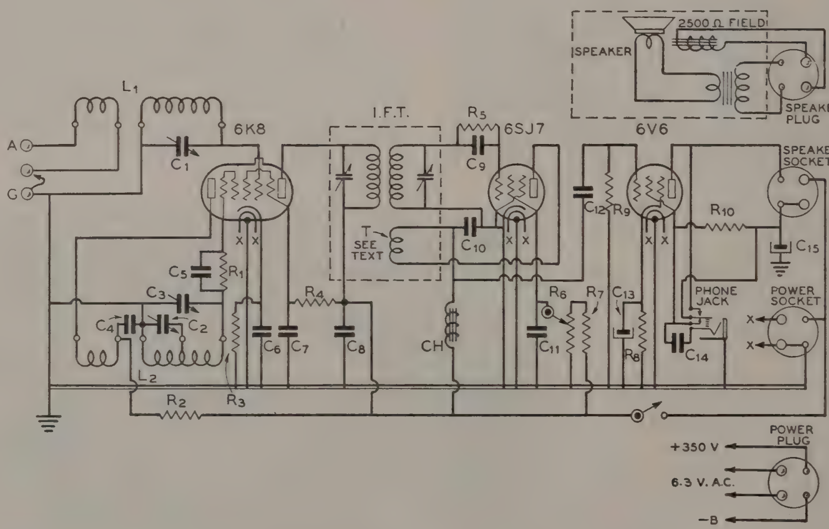


Figure 7.
UNDER-CHASSIS VIEW OF SIMPLE SUPER.

Not much room to spare, but all components fit without crowding. The phone jack is mounted directly on the rear drop of the metal chassis; because of the method of connection, no insulating washers are required.

Figure 8.
WIRING DIAGRAM OF THE 3-TUBE SUPERHET.



C_1, C_2—50-$\mu\mu$fd. midget variable

C_3—140-$\mu\mu$fd. midget variable

C_4, C_8, C_{11}, C_{14}—0.1-μfd. tubular, 400 volts

C_5, C_9—.0001-μfd. tubular, 600 volts

C_6, C_7, C_{12}—.01-μfd. tubular, 600 volts

C_{10}—.001-μfd. tubular, 600 volts

C_{13}—25-μfd. 25 volt electrolytic

C_{15}—4-μfd. 450 volt midget tubular electrolytic

R_1, R_2—50,000 ohms, 1½ watts

R_3—300 ohms, 1 watt

R_4—40,000 ohms, 1½ watts

R_5—5 meg. insulated ½ watt resistor

R_6—100,000-ohm potentiometer

R_7—100,000 ohms, 1½ watts

R_8—400 ohms, 10 watts

R_9—500,000 ohms, 1½ watts

R_{10}—10,000 ohms, 10 watts

IFT—1500 kc. replacement type i.f. trans. (see text for tickler data)

CH—High impedance audio choke, 500 or more hy.

Phone Jack—Two-circuit "filament lighting" type

The only band on which images might be bothersome is the 10-meter band. In most cases, objectionable images can be eliminated without serious loss in signal strength by shifting the h.f. oscillator to the other side by means of the bandset condenser. The receiver will work with the oscillator either *higher or lower* by the intermediate frequency than the received signal. On the higher frequency bands, the bandset condenser tunes over a wide enough band of frequencies so that it hits both sides.

On certain bands the gain and sensitivity are better with the h.f. oscillator on one side of the detector than on the other. Some experimenting with the bandset condenser should be made on those bands where it is possible to hit both the high and low side with the bandset condenser.

COIL TABLE
For Simple Super

160-M. Mixer

58 turns no. 24 enam. close-wound on 1½ in. form, padded with 50-$\mu\mu$fd. midget mica fixed condenser placed inside form; ant. coil 14 turns close-wound at ground end spaced ¼ in. from grid winding

160-M. Osc.—80-M. Mixer

42 turns no. 22 d.c.c. close-wound on 1½ in. form; bandspread tap 20 turns from ground end; tickler 9 turns close-wound, spaced 1/16 in. from main winding

80-M. Osc.—40-M. Mixer

20 turns no. 22 d.c.c. spaced to 1½ in. on 1½ in. form; bandspread tap 12 turns from ground end; tickler 8 turns close-wound, spaced ⅛ in. from main winding

40-M. Osc.—20-M. Mixer

11 turns no. 22 d.c.c. spaced to 1¼ in. on 1½ in. form; bandspread tap 5 turns from ground end; tickler 6 turns close-wound, spaced ⅛ in. from main winding

20-M. Osc.—10-M. Mixer

5½ turns no. 22 d.c.c. spaced to 1 in. on 1¼ in. form; bandspread tap 3 turns from ground end; tickler 4 turns close-wound, spaced 1/16 in. from main winding

10-M. Osc.

3 turns no. 22 d.c.c. spaced to 1 in. on 1¼ in. form; bandspread tap 1½ turns from ground end; tickler 2 turns close-wound, spaced 1/16 in. from main winding

Tickler is always at ground end of main coil. Note that two highest frequency coils are on 1¼ in. forms, rest 1½ in. Tickler polarity must be as shown in diagram to secure oscillation.

Economical 5-Tube Superheterodyne

The sensitivity of the simple superheterodyne just described can be increased by the addition of a tuned r.f. stage ahead of the mixer. The gain and selectivity can be increased by the addition of an i.f. stage. These additions do not add greatly to the total cost, and the improvement in performance makes their incorporation highly desirable. The construction, however, is somewhat more difficult, and should not be attempted as the builder's first effort.

Electrically the receiver is essentially the same as the 3-tube superheterodyne, except for the addition of a 6K7 radio frequency stage and a 6SK7 intermediate frequency amplifier. To minimize the number of tuning controls, the tuning condensers for the r.f. and mixer stages are ganged together.

Mechanical Layout. The r.f. stage is located on the left front corner of the 7 x 11 x 2-inch chassis. The mixer stage is placed at the rear left corner of the chassis, with the shield partition visible in Figure 9 separating it from the r.f. stage. Placing the r.f. and mixer coils toward the edge of the chassis removes them from the proximity of the front-to-back shield, which otherwise might lower the gain obtained in the tuned circuits.

The under-chassis view, Figure 10, shows the location of the two 50-$\mu\mu$fd. ganged condensers used to tune the r.f. and mixer stages. By reversing the usual mounting procedure on these condensers and hanging them stator side down from the chassis, the shafts are brought out at the center of the front drop. A small Isolantite coupling is used to gang the two condensers.

For data on how to wind the tickler turns on the second i.f. transformer, refer to the description given for the 3-tube superheterodyne previously described. The procedure is the same for either receiver. The remarks pertaining to grid hum in the second detector also apply to the 5-tube model.

The receiver is designed for enclosure in a metal cabinet. The cabinet completes the shielding between the r.f. and mixer stages, and prevents oscillation. If a metal cabinet is not used, more elaborate shielding partitions than those illustrated in Figure 9 may be required.

Coils. If the data given in the coil table are followed closely, no trouble should be experienced in getting the r.f. and mixer stages

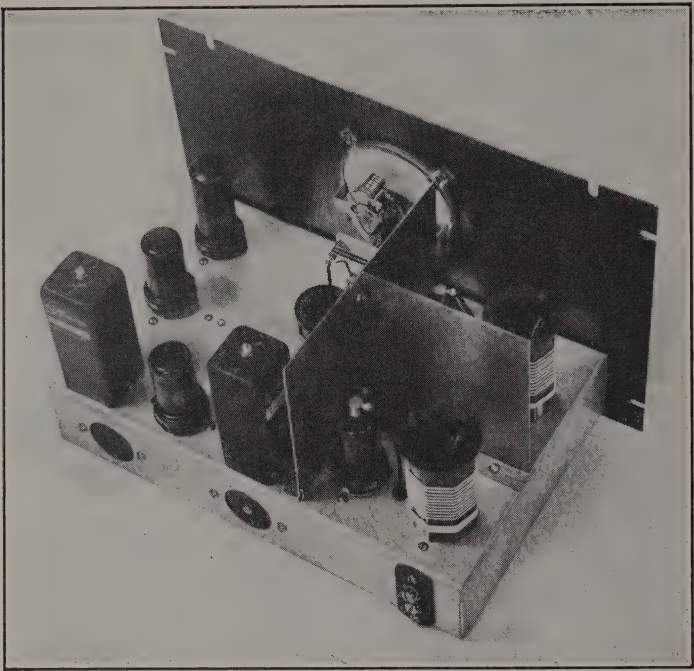


Figure 9.
TOP VIEW OF 5-TUBE SUPERHET.
R.f. stage at the front, mixer at the rear, and the i.f. and audio strung out along the rear and far edge of the chassis. A corner of the oscillator coil may be seen peeking around the front-to-rear shield.

to track accurately. It will be noted that the r.f. and mixer coil secondaries are identical on all bands except 10 meters, where the r.f. stage has one less turn. It is a simple matter to check the tracking. All that is necessary is to loosen the set screws on the coupling between the r.f. and mixer condensers and resonate each condenser separately. By observing the amount of capacity used to resonate each stage near the center of the band in question, it may be determined whether an increase or decrease in the inductance of either coil is necessary.

The oscillator bandspread tap location given in the coil table will give nearly full-dial coverage of each band. Individual constructors who may have different ideas as to the proper amount of bandspread to use may move the taps along the coils to obtain any desired

COIL DATA
For 5-Tube Super

All coils are wound with no. 22 d.c.c. wire

80 Meters

L_1—42 turns on 1½″ dia. form; antenna 7 turns close-wound
L_2—42 turns close-wound on 1½″ dia. form; primary 9 turns close-wound
L_3—20 turns spaced to occupy 1½″ on 1½″ dia. form, tapped 15 turns from ground; tickler 8 turns close-wound

40 Meters

L_1—21 turns spaced to occupy 1½″ on 1½″ dia. form; antenna 6 turns close-wound
L_2—21 turns spaced to occupy 1½″ on 1½″ dia. form; primary 7 turns close-wound
L_3—10 turns spaced to occupy 1¼″ on 1½″ dia. form, tapped 6½ turns from ground; tickler 6 turns close-wound

20 Meters

L_1—11 turns spaced to occupy 1¼″ on 1½″ dia. form; antenna 4 turns close-wound
L_2—11 turns spaced to occupy 1¼″ on 1½″ dia. form; primary 6 turns close-wound
L_3—6 turns spaced to occupy 1″ on 1¼″ dia. form, tapped 4 turns from ground; tickler 4 turns close-wound

10 Meters

L_1—6 turns spaced to occupy 1″ on 1¼″ dia. form; antenna 4 turns close-wound
L_2—7 turns spaced to occupy 1″ on 1¼″ dia. form; primary 4 turns close-wound
L_3—3 turns spaced to occupy 1″ on 1¼″ dia. form, tapped 2 turns from ground end; tickler 3 turns close-wound

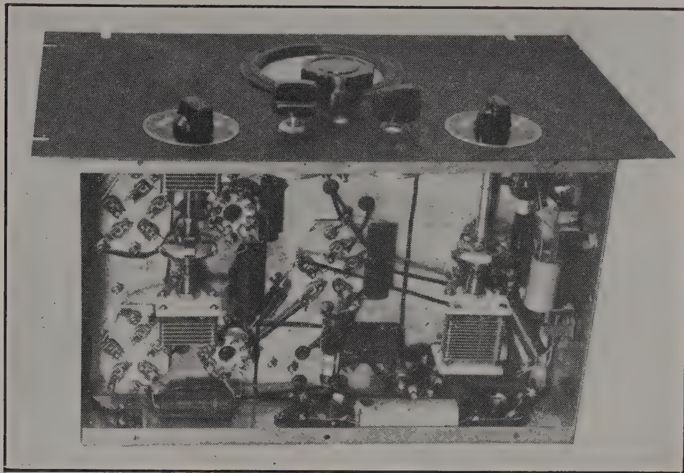


Figure 10.
SHOWING FRONT PANEL AND UNDERSIDE OF CHASSIS.
Most of the "works" are under the chassis. The two ganged r.f. and mixer tuning condensers are visible in this photograph, as is the oscillator bandsetting condenser.

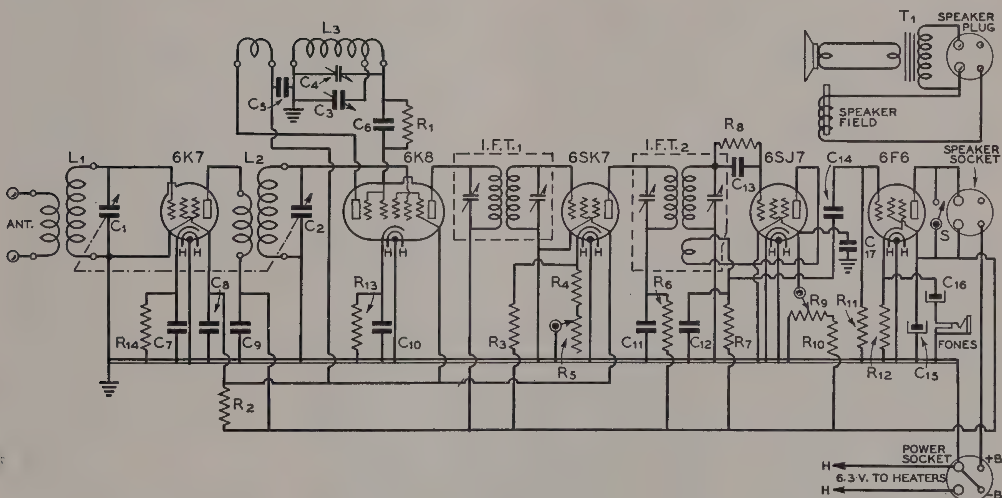


Figure 11.
GENERAL WIRING DIAGRAM OF THE 5-TUBE SUPER.

C_1, C_2—50-$\mu\mu$fd. midget variable
C_3 — 25-$\mu\mu$fd. midget variable
C_4—140-$\mu\mu$fd. midget variable
C_5 — 0.1-μfd. 400-volt tubular
C_6—.0001-μfd. mica
C_7 — .01-μfd. 400-volt tubular
C_8, C_9, C_{10}, C_{11}—0.1-μfd. 400-volt tubular
C_{12}—.0005-μfd. mica
C_{13}—.0001-μfd. mica
C_{14}—.01-μfd. 400-volt tubular
C_{15} — 8-μfd. 450-volt electrolytic
C_{16} — 10-μfd. 25-volt electrolytic
C_{17}—0.1-μfd. 400-volt tubular
Note: Omitted from the diagram was a condenser from the 6SK7 i.f. stage cathode to ground. This condenser should be a .01-μfd. 400-volt unit.
R_1—75,000 ohms, ½ watt
R_2 — 25,000 ohms, 2 watts
R_3 — 60,000 ohms, 1 watt
R_4—300 ohms from stop on R_5
R_5 — 10,000-ohm potentiometer
R_6 — 2000 ohms, ½ watt
R_7—250,000 ohms, ½ watt
R_8—1 megohm, ½ watt
R_9 — 10,000-ohm potentiometer
R_{10}—100,000 ohms, 1 watt
R_{11}—250,000 ohms, ½ watt
R_{12} — 600 ohms, 10 watts
R_{13}, R_{14}—300 ohms, ½ watt
IFT_1—1500-kc. input i.f. transformer
IFT_2—1500-kc. input i.f. transformer (see text for alterations)
S—S.p.s.t. toggle switch
L_1, L_2, L_3—See coil table
T_1 — Pentode output transformer (on speaker chassis)

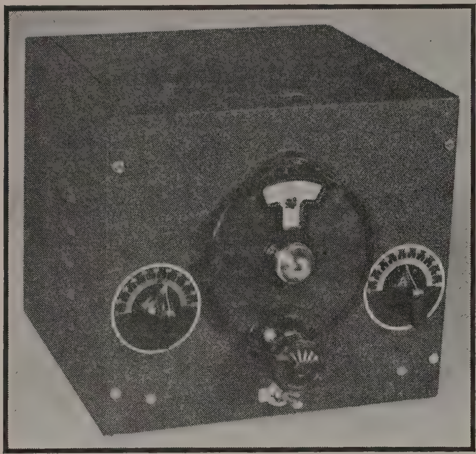


Figure 12.
THE 6K8-6J5 CONVERTER IN ITS CABINET.

This converter may be used ahead of nearly any broadcast receiver to give good high-frequency reception of both 'phone and c.w. signals. The left-hand knob controls the oscillator bandset condenser, while the right-hand knob operates the mixer-section trimmer. The large dial is for bandspread tuning. The switch below the dial controls the b.f.o.

amount. The 20- or 75-meter 'phone bands may be spread across the whole dial, for instance, by moving the taps on the coils for these bands nearer the grounded end. Conversely, any one of the bands may be packed into a few dial divisions by moving the coil tap on that band to the grid end of the coil.

Initial Operation. After the receiver has been connected to a power supply delivering from 250 to 300 volts and a speaker having a field resistance of 1500 to 2500 ohms, the i.f. stage and second detector input circuit should be aligned. This is best accomplished with the aid of a signal generator, operating in the 1500-to-1600 kc. range, coupled loosely to the grid of the mixer. The detector should go into oscillation very smoothly when the regeneration control, R_9, is advanced. If oscillation does not take place it is probable that the tickler is improperly phased, and the tickler connections should be reversed.

After the i.f. amplifier has been aligned, a set of coils should be plugged in and the oscillator bandsetting condenser set to the proper capacity for the coils in use. With a 0–100 scale, with zero at the low capacity end, this setting will be as follows: 10 meters, 35; 20 meters, 80; 40 meters, 60; 80 meters, 60. Next, the r.f. and mixer tuning control should be brought into resonance, and, after the oscillator bandsetting control has been adjusted to center the band on the dial, the receiver is ready for use.

CONVERTERS

Quite often it is desired to adapt a broadcast-band receiver to short-wave reception, or to improve the short-wave reception of an inexpensive "all-wave" receiver. For this purpose, one of the converters described on the following pages may be used. There is no object, of course, in adding a single-tube converter of the type shown in Figures 12 and 13 to a well-designed communications receiver which already has a high-performance converter section plus one or more r.f. stages. On the other hand, a high-gain converter of the type shown in Figures 15 and 16 often can give exceptional results when used with a good communications receiver.

6K8 Converter with B.F.O.

The unit shown in Figures 12 and 13 and diagrammed in Figure 14 is intended for use ahead of "broadcast" or "all-wave broadcast" receivers. Since receivers of this type are not usually equipped with a beat-frequency oscillator for c.w. telegraphy reception, a b.f.o. is included on the converter chassis.

Plug-in coils are used in the converter to cover the amateur bands from 10 to 80 meters with full bandspread on each band. The high-frequency broadcast and commercial code stations between the amateur bands may also be received with the coils shown, by simply adjusting the panel bandsetting condensers to the desired frequency. The use of bandspread allows easy tuning in these commercial ranges, also.

The unit consists essentially of a 6K8 triode-hexode converter tube functioning as a mixer and oscillator with its output on approximately 1600 kc., and a 6J5 b.f.o. also on 1600 kc.

Mechanical Construction. The complete converter is built into a small 7 x 8 x 7½-inch cabinet of standard manufacture. The chassis is also a standard unit and is designed to be used with this particular cabinet.

Three controls and the b.f.o. switch are mounted upon the front panel. The left control is the knob on the bandset control, which consists of a 100-$\mu\mu$fd. variable condenser connected across the oscillator-section coil. The center condenser is the 35-$\mu\mu$fd. midget bandspread condenser, which is operated by the main tuning dial. The right-hand control is the mixer-section tuning condenser, and con-

sists of a 50-$\mu\mu$fd. midget connected directly across the grid coil.

A glance at the photograph of the chassis will reveal the location of most of the components. The oscillator- and mixer-section coil sockets are mounted directly behind their respective bandsetting condensers. Steatite sockets are used for these coils, and also for the 6K8 tube, which is located in the center of the chassis directly behind the bandspread condenser. The output transformer, I.F.T., is located near the left rear corner of the chassis, with the 6J5 and the b.f.o. transformer to its right. The output transformer is made from a standard 1600-kc. iron-core i.f. transformer by simply removing one of the windings and winding about 20 turns of the wire from the discarded coil back around the dowel as close to the remaining winding as possible. One of the leads from the 20-turn winding is grounded, and the other is brought to an auto-type connector at the rear of the chassis. A shielded single-conductor lead should be used to connect the converter to the receiver antenna and ground terminals, the shield acting as the ground connection between the converter and receiver.

Tuning Up. Tuning up the converter is a comparatively simple process, provided the coil table has been followed exactly, and provided a high-gain broadcast receiver is available for the first test. The b.c. set is first tuned to 1600 kc. (or a point close to that frequency where no b.c. or police stations are audible) and the gain turned up until background noise can be heard. The 40-meter coils should be plugged into the converter, as this band is most likely to have plenty of signals day or night.

With the bandset condenser on the oscillator at about half scale, tune the primary on the 1600-kc. output i.f. transformer in the converter at the same time that the mixer-section tuning condenser is being rotated back and forth. A point will be found where the hiss (or perhaps a signal) comes in loudest. The output transformer trimmer should now be adjusted for maximum hiss, after which it should not be touched. Now the oscillator bandsetting condenser should be slowly varied, following with the mixer condenser to keep the background noise at maximum, until the desired signals are found. The mixer trimmer should next be carefully set for maximum signal strength, and the tuning done with the bandspread dial. A single setting of the mixer bandsetting condenser will serve for about half the bandspread condenser tuning range, after which the mixer tuning should again be peaked for maximum signal strength.

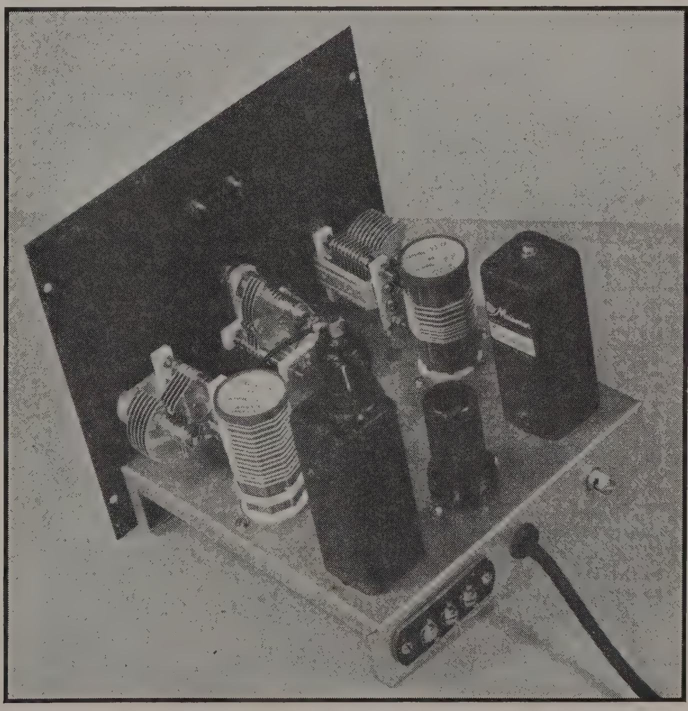


Figure 13.
TOP-REAR VIEW OF THE CONVERTER CHASSIS.
The mixer section is at the left front and the oscillator section at the right front in this view, with the 6K8 between the coils. The b.f.o. and output transformers are at the rear, with the 6J5 between them.

COIL DATA
For 6K8-6J5 Converter

All coils are wound on 1¼″ diameter forms with no. 22 d.c.c. wire

80 Meters

Oscillator—22 turns 1¼″ long, tapped 15 turns from ground; tickler 6 turns close-wound
Mixer—45 turns close-wound; antenna coil 7 turns close-wound

40 Meters

Oscillator—15 turns 1¼″ long, tapped 7 turns from ground; tickler 4 turns close-wound
Mixer — 30 turns close-wound; antenna coil 6 turns close-wound

20 Meters

Oscillator—7 turns 1″ long, tapped 3 turns from ground; tickler 3 turns interwound with grid winding
Mixer—14 turns 1½″ long; antenna coil 5 turns close-wound

10 Meters

Oscillator—3 turns 1¼″ long, tapped 1 turn from ground; tickler 2 turns interwound with grid coil
Mixer—7 turns 1¼″ long; antenna coil 4 turns close-wound

To receive c.w. telegraph signals, the b.f.o. is turned on by means of switch S and the trimmer on the beat-oscillator transformer adjusted for a beat note of suitable pitch. If the beat-oscillator output is too great, the receiver may be blocked, or weak signals may be masked by hiss. Too little beat-oscillator output will not give a "beat" with loud signals. The b.f.o. output may be adjusted by changing the size of R_4; increasing the resistance will increase the output, decreasing it will decrease the output.

High Performance Converter

The converter seen in the photos of Figures 15 and 16 and the diagram of Figure 17 is intended to allow superlative performance on the 10- and 20-meter amateur bands when used in conjunction with a first-class communication receiver. When used with a high-performance communication receiver on the 40- and 80-meter bands, the converter probably will not give any great improvement in sensitivity, since such receivers by themselves are very sensitive in these bands. However, due principally to the limitations of coil switching systems, an improvement in the performance can often be made in the 10- and 20-meter ranges through the use of the converter. With

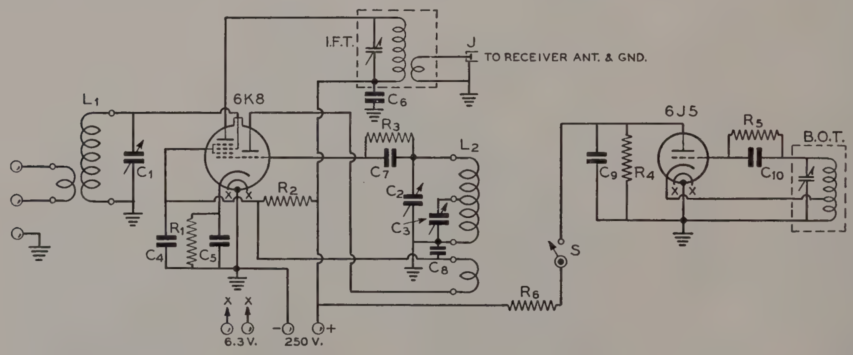


Figure 14.
CONVERTER WIRING DIAGRAM.

C_1—50-$\mu\mu$fd. midget variable
C_2—100-$\mu\mu$fd. midgt variable
C_3—35-$\mu\mu$fd. midget variable
C_4—.005-μfd. mica
C_5, C_6 — .01-μfd. 400-volt tubular
C_7 — .0001 - μfd. mica
C_8—.005-μfd. mica
C_9 — .05-μfd. 400-volt tubular
C_{10} — .0005 - μfd. mica
R_1—300 ohms, ½ watt
R_2—20,000 ohms, 1 watt
R_3 — 50,000 ohms, ½ watt
R_4 — 50,000 ohms, ½ watt
R_5—100,000 ohms, ½ watt
R_6—250,000 ohms, ½ watt
S — S.p.s.t. toggle switch
L_1, L_2—See coil table
I.F.T. — 1600-kc. i.f. transformer, see text for alterations
B.O.T. — 1600-kc. beat-frequency oscillator transformer

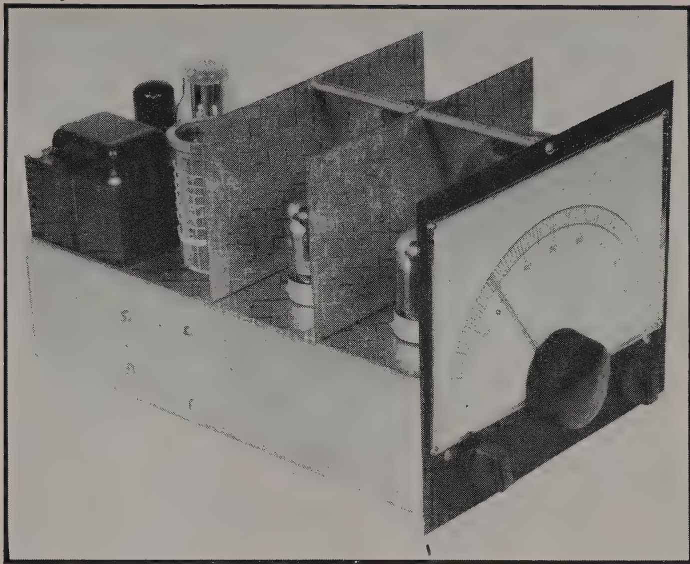


Figure 15.
THE HIGH PERFORMANCE CONVERTER.
The oscillator section is between the panel and first shield, the mixer between the two shields, and the r.f. stage behind the rear shield. The 1232 r.f. stage tube can not be seen in this view; it occupies the same position behind the rear shield as the coils take in the mixer and oscillator sections. A manufactured dial similar to the home-built one shown is now available.

modest receivers in the lower priced brackets (especially those not having an r.f. stage) the converter will also provide improved performance on 40 and 80 meters. For this reason, data on 40- and 80-meter coils are included in the coil table; the converter might just as well be used on every band where its use results in improved performance.

While the heater and plate current requirements are not great, they are somewhat more than can be taken safely from the plate supply of many commercial receivers. Hence, a small power pack has been made an integral part of the unit. To prevent drift as a result of heating of the oscillator components, the oscillator is placed to the front of the chassis and the power pack to the extreme rear. This also gives a positive drive on the oscillator tuning condenser, as there is no flexible coupling between dial and oscillator tuning condenser to permit backlash. To stabilize the oscillator against frequency changes due to plate voltage changes (as a result of line voltage changes), a voltage regulator tube is used.

To provide maximum conversion gain in the mixer, grid leak bias and control grid injection are employed. This, in conjunction with the high gain tubes used both in the r.f. and mixer stages, gives a high potential overall gain. The full potential gain is closely approached as a result of fairly low-C low-loss tank circuits, difficult to obtain on high frequency bands with all-band bandswitching but easily obtained with plug-in coils when proper mechanical layout is employed. The oscillator is made high-C for the sake of stability, as it provides more than sufficient excitation when control grid injection is employed.

For maximum signal-to-noise ratio, the r.f. stage is run "wide open" on weak signals. However, the converter has so much gain that the receiver with which it is used may be blocked on loud local signals. Hence, an r.f. gain control (R_1) is provided. This is normally left full on, and backed off only when a signal is so loud that there is blocking.

Construction. The converter is constructed on a 7 x 12 x 3-inch chassis, with a front panel 7 inches high by 8 inches wide. The two stage shields are standard $5\frac{1}{2}$ x 7-inch manufactured items, cut down to a height of 4 inches.

Under the chassis, a partition is placed $5\frac{1}{4}$ inches from the rear drop. It is cut from a piece of 20 gauge sheet metal. This partition is $2\frac{3}{4}$ inches high and serves both as a shield between the r.f. and mixer stages, and as a mounting support for the antenna switch and the gain control, both of which are operated from the front panel by means of shaft extensions. No shielding is employed between the oscillator and mixer condensers, as it is unnecessary.

All three tuning condensers are bolted directly to the underside of the main chassis, and are connected by means of flexible shaft couplings. The arrangement of the balance of the components should be clear from inspection of the illustrations.

Output Transformer. The output transformer, IFT, is a modified 1500-kc. i.f. trans-

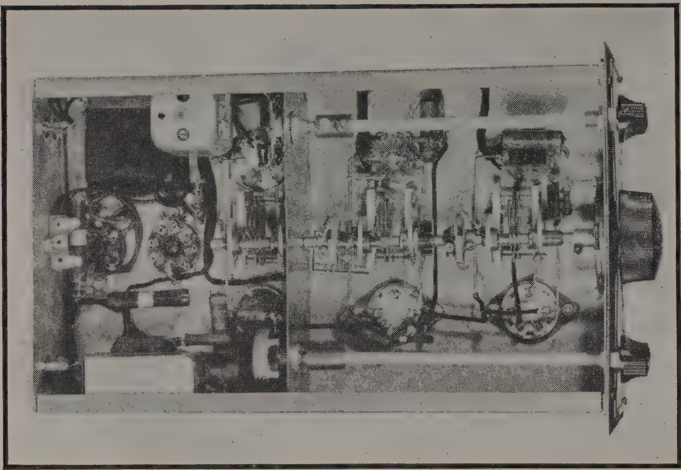


Figure 16.
UNDER-CHASSIS VIEW OF THE CONVERTER.
Note the shield partition between the r.f. and mixer-oscillator sections. The gain control and antenna switch are mounted on this shield.

former. The secondary winding is removed, and in its place are wound 15 turns of no. 22 d.c.c. wire as close to the primary coil as possible. To permit mounting of the transformer under the chassis, its shield can is cut off to a height of about 2 inches.

The Dial. The dial seen in the photographs is constructed around a planetary reduction unit supplied as part of a manufactured vernier dial. Since the dial was built, however, the manufacturers of the reduction unit have themselves introduced a similar dial, which can be used in place of the home-made one shown. It will be necessary to cut a small slot in the top front of the chassis to take the top portion of the dial reduction unit.

Injection. The control-grid injection employed provides high conversion gain, and is

COIL DATA
High-Performance Converter

All coils are wound with no. 22 d.c.c. wire

28 Mc.

Oscillator—3 turns spaced to occupy 1¼″ on 1½″ dia. form; bandspread tap 1¼ turns from ground; cathode tap 1 turn from ground. C_6 is 50-μμfd. air trimmer.

Mixer—6 turns spaced to occupy 1½″ on 1½″ dia. form, tapped 1⅞ turns from ground; primary 3 turns close-wound; C_5 is 12-μμfd. ceramic-based mica trimmer.

R.F. Stage—Same as mixer.

14 Mc.

Oscillator—9 turns spaced to occupy 1¼″ on 1½″ dia. form; bandspread tap 3 turns from ground; cathode tap 2½ turns from ground. C_6 is 75-μμfd. air trimmer.

Mixer—12 turns spaced to occupy 1½″ on 1½″ dia. form, tapped 3 turns from ground; primary 6 turns close-wound; C_5 is 12-μμfd. ceramic-based mica.

R.F. Stage—Same as mixer except primary has 5 turns close-wound.

7 Mc.

Oscillator—18 turns spaced to occupy 1½″ on 1½″ dia. form; bandspread tap 6 turns from ground, cathode tap 5½ turns from ground. C_6 is 75-μμfd. air trimmer.

Mixer—23 turns spaced to occupy 1½″ on 1½″ dia. form, tapped 8 turns from ground; primary 12 turns close-wound; C_5 is 35-μμfd. ceramic-based mica trimmer.

R.F. Stage—Same as mixer except primary has 8 turns close-wound.

3.5 Mc.

Oscillator—21 turns close-wound on 1½″ dia. form; bandspread tap 13 turns from ground, cathode tap 7 turns from ground. C_6 is 75-μμfd. mica trimmer.

Mixer—40 turns close-wound on 1½″ dia. form, tapped 26 turns from ground; primary 14 turns close-wound. C_5 is 35-μμfd. mica trimmer.

R.F. Stage—Same as mixer except primary 10 turns close-wound.

All primary windings are placed at ground end of grid windings, and spaced approximately ¼″ therefrom.

not especially critical as to coupling. However, if too little coupling is used, there will be a reduction in conversion gain, and if too much coupling is used, there will likewise be a reduction in gain together with "pulling" between the mixer and h.f. oscillator when aligning the coils. Two pieces of pushback hookup wire, twisted together for about ½ inch, will be found to provide about the right amount of coupling if the exact mechanical layout illustrated is followed. Because of the stray capacity coupling present, very little additional coupling capacity will be required.

The type of control-grid injection used in this converter has two advantages: it allows the mixer cathode to be grounded directly without an intervening resistor and by-pass condenser, and it allows the mixer bias to automatically adjust itself to the proper amount for correct operation, even though the oscillator output varies widely. Since the mixer bias is provided by a grid leak (R_5) the circuit does have the disadvantage, however, that when the mixer is not receiving excitation from the oscillator there is no bias on the mixer, and it may draw excessive plate and screen current. For this reason, the oscillator coil should never be changed without first turning off the plate supply.

The Coils. All coils are wound on standard 1½-inch 5-prong forms with no. 22 d.c.c. wire. The padding condensers are mounted

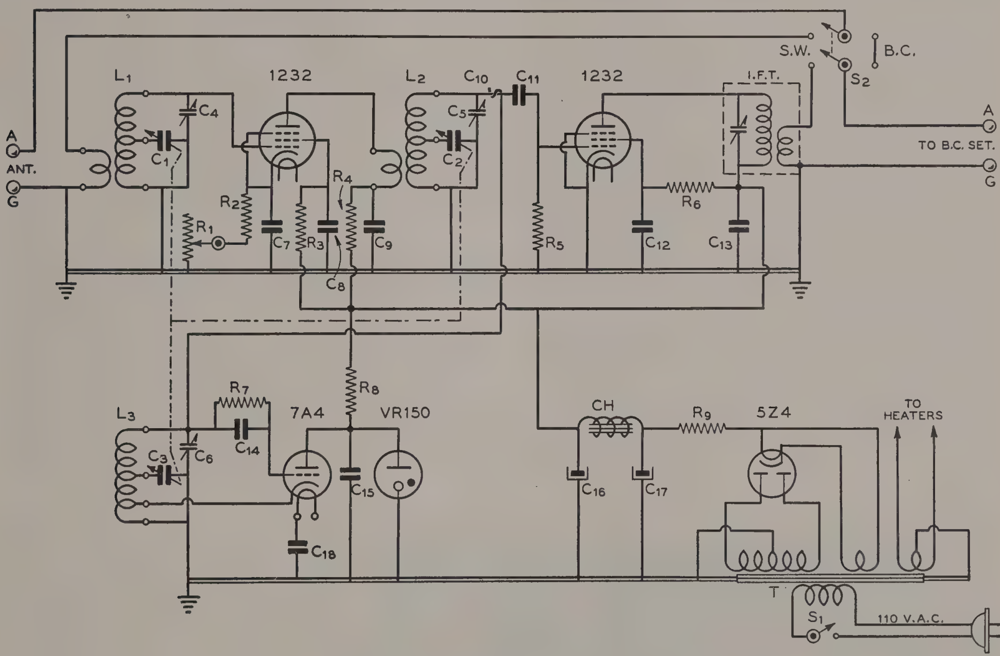


Figure 17.
WIRING DIAGRAM OF THE CONVERTER.

C_1, C_2, C_3—35-μμfd. midget variable, straight line capacity (semi-circular)
C_4, C_5, C_6—Mounted inside coils, see coil table
C_7, C_8, C_9—.01-μfd. 400-volt tubular
C_{10}—Injection coupling; see text
C_{11} — .0001 - μfd. mica
C_{12}, C_{13} — .01-μfd. 400-volt tubular
C_{14} — .0001 - μfd. mica
C_{15} — .01-μfd. 400-volt tubular
C_{16}, C_{17} — 8 - μfd. 450-volt electrolytic
C_{18} — .01-μfd. 400-volt tubular
R_1 — 10,000 - ohm potentiometer
R_2—250 ohms, ½ watt
R_3 — 75,000 ohms, ½ watt
R_4—2000 ohms, ½ watt
R_5—5 megohms, ½ watt
R_6 — 75,000 ohms, ½ watt
R_7 — 50,000 ohms, ½ watt
R_8—5000 ohms, 10 watts
R_9—2000 ohms, 10 watts
S_1—S.p.s.t. (on R_1)
S_2 — D.p.d.t. tap switch
IFT — Modified 1500-kc.i.f. transformer; see text
T—700 v. c.t., 70 ma.; 5 v., 3 a.; 6.3 v. c.t., 2.5 a.
CH—10 hy., 40 ma.
L_1, L_2, L_3—See coil table

inside the coil forms, ceramic compression type being used for the r.f. and detector, and air tuned type for the oscillator coils. If the coil specifications are followed exactly, no difficulty should be experienced in getting the coils to track.

The oscillator is operated on the "high side" (intermediate frequency higher than signal frequency) on all bands. Getting the coils to track is a simple procedure if the coil data is followed closely.

To align the coils, simply insert the proper coils for a band, adjust the oscillator padder (bandset condenser) to center the band on the dial (making sure the oscillator is on the proper side of the signal frequency), and then, after tuning in a signal near the center of the dial, peak up the mixer and r.f. trimmers for maximum signal.

VARIABLE-SELECTIVITY CRYSTAL FILTER

The variable-selectivity crystal filter unit pictured in Figures 18 and 20 and diagrammed in Figure 19 may be built for about $12, including the crystal and the output coupling tube. When used with a small, inexpensive communications receiver, the unit gives the owner selectivity comparable with that of the most expensive sets.

The theory of the crystal filter operation is discussed in Chapter 4, so the following description will concern only the mechanical details and operation of the unit.

Construction. The complete filter, including the 6SK7 output coupling tube, is contained in a 3 x 4 x 5-inch metal box. As the photos show, the unit is constructed on an aluminum chassis which mounts vertically to one of the 4-inch sides of the box. The chassis is $2\frac{7}{8}$ inches wide, $1\frac{3}{4}$ inches high, and $3\frac{7}{8}$ inches long. A shield partition divides the underside of the chassis into two separate sections. On one side of the shield is the wiring associated with the input circuit up as far as the crystal and phasing condenser, while on the other side is the output circuit and the remainder of the components. It is absolutely essential that this shield be employed, since coupling around the crystal between the input and output circuits in any manner whatsoever will completely ruin the selectivity characteristic of the unit.

A single standard i.f. transformer, which may be of either the "input" or "interstage" type, serves to provide parts for both T_1 and T_2. The transformer should be removed from its shield can and, after disconnecting the leads from the bottom coil to its trimmer, saw through the dowel about 3/16 inch below the cardboard disc under the top winding. The sawed-off bottom section of the transformer becomes the coil for T_2 when mounted to the shield partition by means of a short brass wood screw.

The secondary of T_1 is made by winding about 100 turns (the exact number is not critical) of small wire in a slot formed between the cardboard disc originally on the transformer and another circular piece of cardboard which is glued across the bottom of the dowel. If the dowel was cut off 3/16 inch below the original disc, this will make a winding slot 3/16 inch wide and about $\frac{1}{2}$ inch deep in which

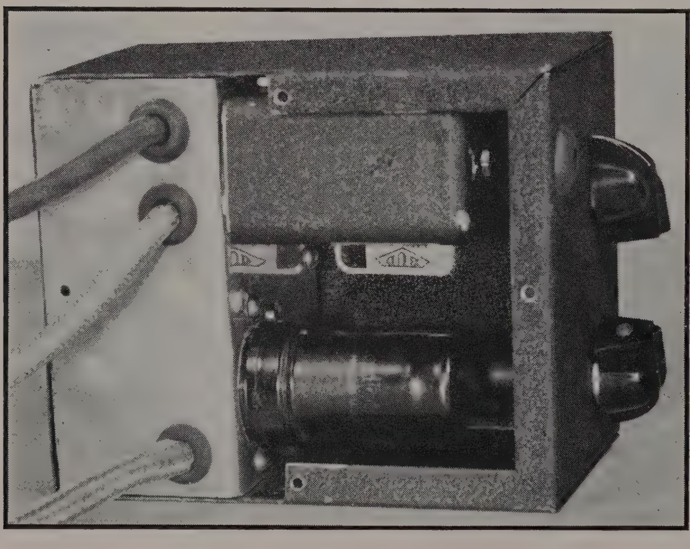


Figure 18.
ACCESSORY CRYSTAL FILTER.
Looking into the crystal with the left side plate removed. The tube and the input transformer are in the foreground in this view. The shafts to the phasing (top) and selectivity-control (bottom) condensers are hidden from view, as is the crystal, which is behind the input transformer. The input transformer primary is trimmed to resonance through the grommet-filled hole at the upper left of the panel. Note the large shielded leads for input and output connections.

to put the secondary winding. In the unit shown, this winding was made with some wire from an old i.f. winding, which happened to be handy at the moment. Any small silk- or cotton-covered wire—no. 30 or thereabouts—will do, however. The ends of the winding may be secured to the unused trimmer terminals, after which the trimmer plates may be nipped off with a pair of diagonals. Enough of the trimmer plates should be left to keep the terminal tabs from pulling through the holes in the ceramic mounting plate.

Before the input transformer is reassembled, its shield can should be cut down to a height of $2\frac{1}{2}$ inches, and a pair of spade bolts secured to the bottom edge by screws or rivets to allow it to be firmly mounted to the chassis.

On the output side of the under-chassis shield are mounted the output winding, T_2, the selectivity control, C_6, the fixed condenser across T_2, and the output coupling tube with its associated resistors and by-pass condensers. A zero-temperature-coefficient padding condenser (C_5) is used across the output winding for padding purposes.

A small 25-$\mu\mu$fd. variable condenser is used for C_6, the selectivity control. When the output circuit is tuned to resonance by means of this condenser, the selectivity is at a minimum. As the circuit is detuned from resonance the selectivity increases.

Wiring. No particular precautions need be observed in the wiring of the unit except the very important one of keeping the input and output circuits well separated and shielded from each other at all points. The only other trouble which might occur would be oscillation in the output stage. Any possibility of oscillation may be obviated by locating the screen and cathode by-pass condensers for the 6SK7 so that they lie across the socket between the grid and plate terminals.

The shaft couplings to C_3 and C_6 are made from a piece of an inexpensive bakelite trimming tool. The hexagonal hole in the tool is small enough so that it makes a tight fit when placed over the nut on the end of the condenser shaft. A standard $\frac{1}{4}$-inch bakelite shaft is run into the other end of the coupling and held in place by means of a 4-40 machine screw in a hole tapped through the coupling.

Leads to and from the filter are made from low-capacity shielded wire. The leads should not be any longer than absolutely necessary, since the additional capacity accompanying the superfluous length may be enough to prevent obtaining resonance in the input and output tuned circuits. A 3-wire cable is used to supply filament and positive B power to the unit, the negative connection being made through the shield on the input and output leads.

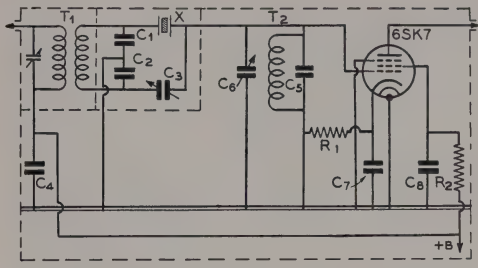


Figure 19.
CRYSTAL FILTER DIAGRAM.

C_1, C_2,—.0001-μfd. mica

C_3—25-$\mu\mu$fd. midget variable, one rotor and two stator plates removed. A corner of one of the rotor plates should be bent over to short out the condenser at the maximum capacity setting.

C_4 — .01-μfd. 400-volt tubular

C_5—100-$\mu\mu$fd. zero-temperature-coefficient fixed padder

C_6—25-$\mu\mu$fd. midget variable

C_7, C_8 — .01-μfd. 400-volt tubular

R_1 — 500 ohms, ½ watt

R_2—100,000 ohms, ½ watt. This resistor may be eliminated and the screen run directly to positive lead if the receiver has a 100-volt plate supply.

T_1, T_2—See text

X — Filter crystal, 450-500 kc. (To match receiver intermediate frequency)

In order to fit the chassis into the cabinet, it is necessary to do some tailoring on the cabinet. The lip along one side of what is to be the back of the cabinet should be removed by means of tin snips or a hack saw and enough of the lips on the same side of the top and bottom removed to allow the chassis to be slid into the cabinet. The chassis is held in the cabinet by means of a 2-inch-long 6-32 machine screw through the rear of the box and the chassis.

Operation. The filter is intended to be used with receivers having one stage of 450- to 500-kc. i.f. amplification. With receivers having more than one stage, trouble may be encountered with overall oscillation of the i.f. amplifier when the filter unit is added, since the filter contributes some gain. A test will determine whether the filter can be used ahead of two i.f. stages in any particular receiver. If the gain is excessive and the receiver oscillates, R_1 can be raised to about 5000 to 10,000 ohms.

To place the filter in use the input lead is connected to the plate of the receiver mixer tube, and the output lead connected to the

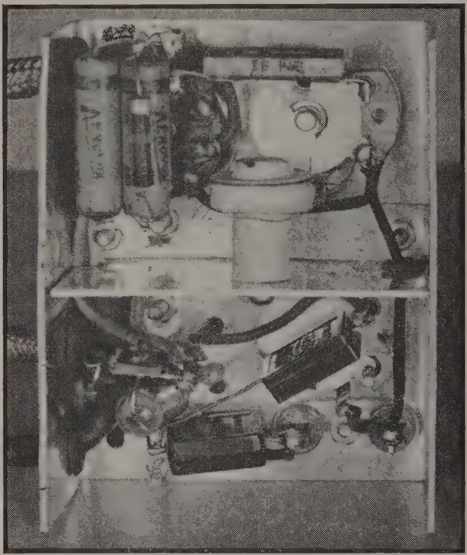


Figure 20.
UNDER THE CRYSTAL FILTER CHASSIS.
Note the shield between the input and output sections of the filter. The input section is at the bottom, the output circuits at the top, in this view. The bypass condensers for the 6SK7 are placed across the tube socket to help prevent oscillation from capacity coupling between the grid and plate.

"plate" lead from the first i.f. transformer in the receiver. The latter lead is, of course, disconnected from the mixer plate. Filament power for the filter unit may be obtained from any convenient point in the receiver, while B power should be taken from the common B positive lead in the receiver.

Before trying to tune up the filter, it is wise to check to see if there is any coupling in the receiver itself which might by-pass signals around the filter. This test may be made by connecting the filter into the receiver in the manner described above, and then removing the 6SK7 output coupling tube from the filter. With the tube removed, there should be no signals passing around the filter, and the receiver should be dead. Any external coupling which allows signals to pass around the filter should be eliminated, as it will reduce the maximum selectivity which may be obtained.

After checking for coupling around the filter and eliminating any that may be found, the tube in the filter may be replaced and the tuning process started. As the receiver intermediate frequency should not be greatly different than that of the crystal, the preliminary tuning may be done by listening to the background noise. Simply adjust the primary trimmer on T_1, condenser C_6, and the trimmer across the receiver i.f. transformer following the filter for maximum noise. The proper setting of the phasing condenser, C_3, will be with the plates about one-third meshed, if the components shown in the original version are used.

Next, a steady signal should be picked up on the receiver and, with C_6 detuned somewhat from the "maximum noise" position, the i.f. amplifier in the receiver is peaked to the crystal. The primary trimmer on T_1 should now be checked again to make sure it is accurately tuned to resonance.

With the above adjustments made, the selectivity control, C_6, should be detuned from resonance as far as possible (maximum selectivity), which should result in a pronounced ringing sound in the phones or speaker, and final trimming adjustments made with the signal accurately tuned in. A check on the operation of the filter may be made by setting the selectivity at maximum, switching on the receiver's beat oscillator, and accurately tuning in a c.w. signal—the correct way to tune is so that the pitch of the beat note is identical with that of the background noise. If the filter is working properly, the c.w. signal will stay at approximately the same strength as the selectivity control is turned toward the broad position, while the background noise and interfering signals will increase greatly in strength.

'Phone Operation. For 'phone reception the filter may be adjusted for as little or as much selectivity as the situation requires. When QRM is not bad, the filter may be cut out entirely by running the phasing condenser to maximum capacity, so that the bent corner of the rotor plate shorts it out. With slight QRM, the filter may be cut in and the selectivity control set at the broadest position, which results in a great reduction in heterodynes, while not greatly reducing the high-frequency speech components. Signals covered by QRM may be made readable by advancing the selectivity control toward maximum. Naturally, the sidebands of the desired signal are clipped by increasing the selectivity, thus reducing the highs, but signals may often be fully understood which would be completely lost without the filter.

One important use of the crystal filter is to eliminate 'phone heterodynes. This is done by first accurately tuning in the desired signal and then carefully adjusting the phasing condenser to drop the heterodyning signal into the crystal rejection notch. The ability of the crystal to eliminate heterodynes in this way depends upon the setting of the selectivity control. The closer the desired and interfering signals are

together (the lower the heterodyne pitch) the more the selectivity must be increased to allow the rejection to be effective. At the maximum selectivity setting the filter will allow heterodynes down to a few hundred cycles to be eliminated.

HIGH-SELECTIVITY I.F. STAGE

With a near-ideal i.f. amplifier having a very narrow, steep-sided curve, it is possible to reject an S9 signal and pull in an S5 signal so closely adjacent that when the receiver is tuned half way between, a heterodyne is audible. When the unit shown in Figure 21 is used in conjunction with a superhet having a reasonably sharp two-stage i.f. amplifier, it is possible to do just that, and when used with a set having one i.f. stage it will do nearly as well. It is possible to tune right up to extremely strong local signals with no interference, unless of course they are splattering or have excessive harmonic distortion.

The device is essentially a "gainless amplifier" stage that can be attached quite easily to any superhet having a 455- or 465-kc. i.f. channel. In the interest of obtaining maximum selectivity, the stage will show little or no gain when correctly adjusted, as it is assumed that the receiver to which it is attached had enough i.f. gain to begin with.

The two transformers are the sharpest inexpensive 455–465 kc. units generally available, being designated by the manufacturer as "interstage" type. The rated gain for one of these transformers is only about half that of a standard input transformer, due to the fact that the coupling between windings is somewhat less than critical; but the selectivity is greater.

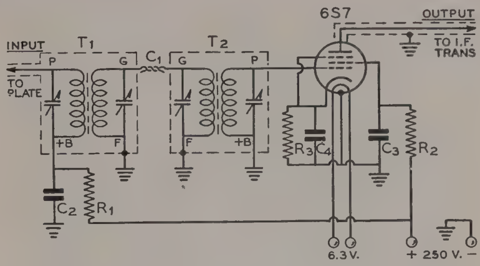


Figure 22.
DIAGRAM OF THE OUTBOARD I.F. STAGE.

T_1, T_2—High Q, iron-core 455-465 kc. interstage i.f. transformers having less than critical coupling. See text.
C_1—I.f. transformer leads twisted together about 1½ inches. See text.
C_2, C_3, C_4—.05-μfd. tubular
R_1—2000 ohms, ½ watt
R_2—75,000 ohms, 1 watt
R_3—500 ohms, ½ watt

To obtain the greatest possible amount of selectivity in one stage, two of these transformers are used in cascade. Coupling between transformers is accomplished by a very small coupling capacity (obtained by twisting together the wire leads for about 1 or 1½ inches).

As the unit is designed to run from the power pack of the regular receiver, a 150-ma.

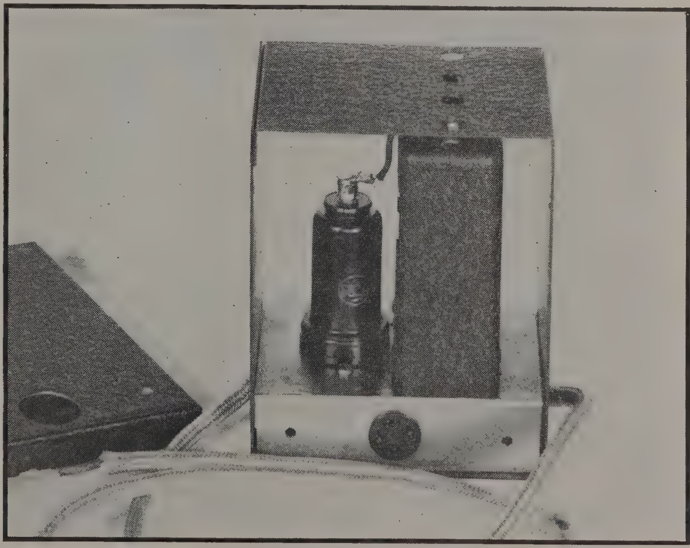


Figure 21.
OUTBOARD SELECTIVE I.F. STAGE.
By adding this unit, which uses a single tube and two inexpensive i.f. transformers, the selectivity of some receivers can be greatly improved.

heater type tube is used to minimize the additional drain placed on the power transformer. As the plate current drawn by the tube is only a few ma., there is no need to fear there being an unduly heavy load placed on the power pack. The unit is quite compact, and will even fit inside the cabinet of some receivers. Where it will not, it may be attached to the outside rear of the receiver cabinet by means of brackets.

As there are so few components in the device, the construction is not a difficult problem. Almost any small shield can, sufficiently large to house the two transformers and the tube with enough room to spare for the resistors and by-pass condensers, will serve the purpose. The shield can shown in the photograph measures 4 x 5 x 3 inches, and has two removable sides. Before the unit is installed in the shield can, four holes must be drilled in the top of the can to permit the transformer trimmers to be adjusted after the unit is permanently installed.

For connecting the unit, a 4-wire cable, with appropriate miniature plug and jack if desired, is used for power connections. Both heater leads are brought through the cable, as in some sets one side of the heater winding is not grounded as is the usual practice. Low capacity shielded wire should be used for the input and output leads. The type of wire commonly employed for making connection to auto radio antennas will be found satisfactory. If the wire has too much shunt capacity per unit length, or if the leads cannot be kept shorter than about 18 inches, trouble may be experienced in getting some of the trimmers in the receiver to "hit." In extreme cases it may be necessary to realign the whole i.f. system on a slightly lower frequency. However, in most receivers the same intermediate frequency may be used if the input and output leads are kept reasonably short. The B plus may be obtained from the screen of the output tube in the receiver, or almost any point at approximately 250 volts potential which does not have excessive series resistance.

Before connecting the unit to the receiver, observe the strength of a local broadcast station (one that is not so loud that it blocks the receiver) on the R meter. If the set does not have an R meter or "eye," a 10-ma. meter may be placed in the plate or cathode lead of the i.f. tube (or one of the i.f. tubes) in the receiver. This will give a reference gain indication. If the set does not have an R indicator, the milliammeter can be used later for realigning the set when the outboard amplifier is added.

The coupling capacity, C, consists simply of the two "grid" wires from the transformers twisted together for about 1 or $1\frac{1}{2}$ inches. The idea is to use the loosest possible coupling at this point which will not result in an actual loss in gain. The wires should be twisted together more tightly or untwisted until the receiver gain is the same as it was before the outboard amplifier was added. A further increase in coupling capacity will result in a voltage gain through the outboard amplifier, but the selectivity will be less (a "double peak" selectivity curve occurring in some cases) and some of the trimmer adjustments will tend to interlock, the adjustment of one trimmer affecting the others.

Instead of twisting the two lead wires together, a small 3–30 $\mu\mu$fd. mica trimmer may be substituted if desired. About 6 or 8 $\mu\mu$fd. will be found optimum; use the least capacity which allows sufficient gain.

The best place to connect the outboard amplifier to the receiver is probably at the plate of the i.f. stage (or the first i.f. stage, if the receiver has more than one). Unsolder the lead from the plate to the primary of the i.f. transformer, and connect the shielded "input" lead to the plate of the i.f. tube and the lead marked "output" to the loose wire on the primary of the i.f. transformer. It will be best to use shielded wire right up to the receiver connections, and not just up to where the input and output wires enter the receiver cabinet.

HIGH GAIN 5-BAND PRESELECTOR

If a superheterodyne has less than two stages of preselection, its performance often can be greatly improved by the addition of this high gain preselector. The improvement in image ratio and signal-to-noise ratio will be most noticeable on the higher frequency bands, and will be especially noticeable if the receiver itself has no r.f. stage at all.

The preselector uses a type 1851 pentode. This tube has a low noise level and extremely high transconductance. In fact, it is necessary to tap the plate of the tube down from the "hot" end of the tuned plate coil in order to avoid oscillation.

The tuned plate circuit is link-coupled to the input terminals of the receiver to which the preselector is to be attached. The coupling link is of the coaxial type, made of flexible shielded conductor. The use of a tuned output circuit and an efficient coupling system makes this preselector greatly superior in performance to the simpler, more common type of 1-stage preselector in which the plate of the preselector tube is capacitively coupled to the antenna post of the receiver.

The preselector is moderately regenerative; in fact, it will tend to oscillate unless the input circuit is rather heavily coupled to an antenna.

The 1851 has a very low input resistance, especially on 10 meters. For this reason the grid is tapped down on the input coil, being connected approximately to the center of the coil. This reduces the grid loading to one-quarter without reducing the input voltage, due to the higher impedance obtained with the tapped arrangement.

Tapping the grid and plate leads down on their respective coils effectively reduces the minimum shunt capacities, thus allowing a greater tuning range with a given tuning condenser. With the 50-$\mu\mu$fd. tuning condensers illustrated, approximately a 2–1 range in frequency is possible with each set of coils. This gives practically continuous coverage of the short-wave spectrum with the coils listed in the coil table. The coils cover the following ranges: 1.7 to 3.5 Mc., 3 to 6 Mc., 6.5 to 11 Mc., 10 to 19 Mc., and 18 to 33 Mc. Thus, the preselector can be used effectively with communication receivers of the continuous coverage all-wave type.

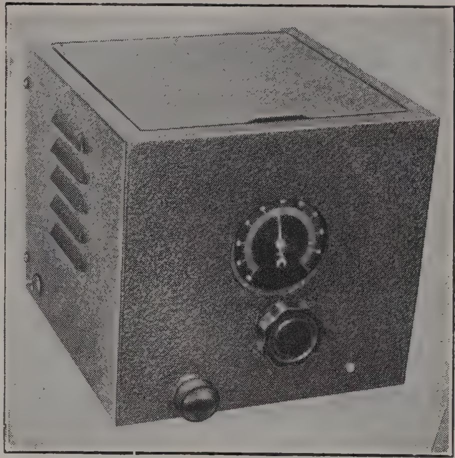


Figure 23.
5-BAND HIGH GAIN PRESELECTOR.
This high gain preselector uses an 1851 tube, tuned output circuit, and moderate regeneration. It makes a worthwhile addition to any receiver having less than 2 r.f. stages.

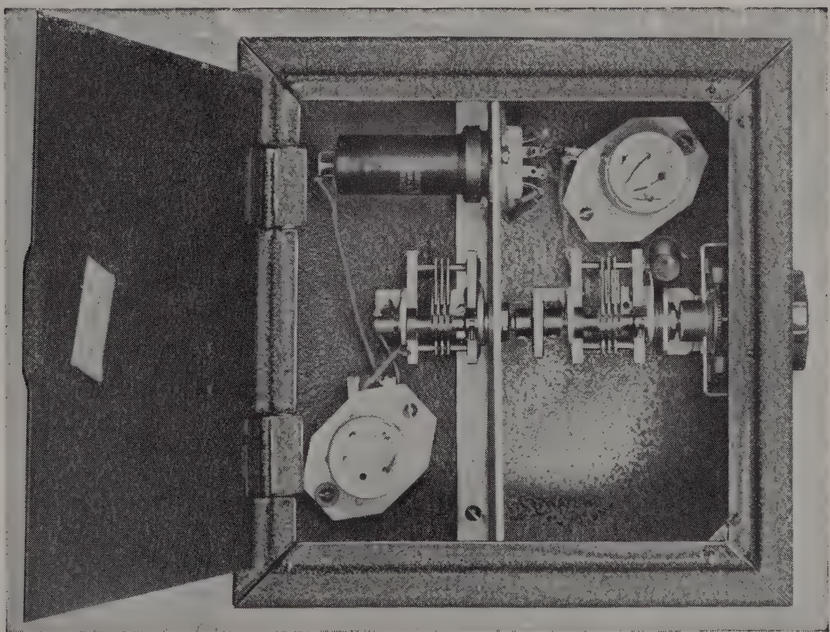


Figure 24.
LOOKING DOWN INTO THE 1851 HIGH GAIN PRESELECTOR.
An aluminum partition shields the input from the output circuit, and serves as a support for the tube and rear tuning condenser.

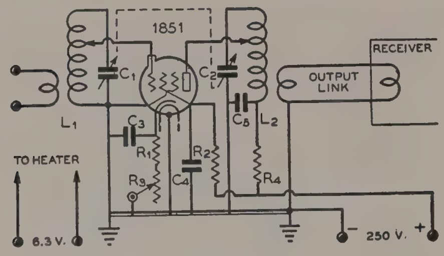


Figure 25.
SCHEMATIC CIRCUIT OF THE 1851 PRESELECTOR.

C_1, C_2 — 50-$\mu\mu$fd. midget variable
C_3 — 0.1-μfd. 400-volt tubular
C_4, C_5 — .01-μfd. 400-volt tubular
R_1 — 200 ohms, 1 watt
R_2 — 50,000 ohms, ½ watt
R_3 — 10,000 - ohm potentiometer
R_4—5000 ohms, 1 watt
Coils—See coil table

If oscillation is troublesome even when tight antenna coupling is used, the plate coil can be tapped a little farther down towards the ground (B plus) end.

If desired, a 6J7 or 6K7 can be used in place of the 1851. If one of these tubes is used, both grid and plate should be connected directly to the "hot" ends of their respective coils, instead of to the center. R_2 should be increased to 100,000 ohms. The gain will not be quite as high as with an 1851, and the tuning range will be reduced slightly. The latter can be offset by using 75-$\mu\mu$fd. tuning condensers instead of 50-$\mu\mu$fd. condensers.

Tracking can be checked by rotating the rear tuning condenser separately while listening to a station and watching the R meter.

Construction. The unit is built in a 7 x 7 x 7-inch cabinet and chassis. A $6\frac{1}{4}$ x $5\frac{1}{4}$-inch aluminum partition with a $\frac{1}{2}$-inch lip to permit fastening to the chassis, as illustrated in Figure 24, shields the input from the output circuits. The rear tuning condenser is mounted on this partition and driven from the front condenser by means of an insulated coupling.

For maximum gain on the higher frequency range, tuning condensers, sockets, and coil forms should have ceramic insulation.

Most receivers will stand a slight additional drain on the plate and filament supplies without overheating. For this reason, the preselector voltages may be robbed from the receiver with which it is to be used. If the receiver power supply already runs quite hot, indicating that it is being overloaded, a separate power supply for the preselector is to be preferred.

COIL DATA
1851 Preselector

160 Meters

Grid—80 turns of no. 26 enam. close-wound on 1½" dia. form; tapped 20 turns from ground; primary 12 turns
Plate—Same as grid; secondary 3 turns

80 Meters

Grid—44 turns no. 22 d.c.c. close-wound on 1½" dia. form; tapped 15 turns from ground; primary 8 turns
Plate—Same as grid; secondary 3 turns

40 Meters

Grid—24 turns of no. 22 d.c.c. spaced to occupy 1½" on 1½" dia. form; tapped 10 turns from ground; primary 5 turns
Plate—Same as grid except tap 12 turns from ground; secondary 3 turns

20 Meters

Grid—15 turns of no. 20 d.c.c. spaced to occupy 1" on 1⅛" dia. form; tapped at center; primary 3 turns
Plate—Same as grid; secondary 2 turns

10 Meters

Grid—8 turns of no. 20 d.c.c. spaced to occupy 1" on 1⅛" dia. form; tapped at center; primary 2 turns
Plate—Same as grid; secondary 2 turns

CHAPTER SEVEN

Transmitter Theory

THE general function of a transmitter is to generate a signal of a desired frequency and to modulate this signal in accordance with the intelligence to be transmitted. The radio frequency energy from the transmitter is most commonly carried by a *transmission line* to a radiating system or *antenna* from whence the intelligence-carrying energy is radiated into space. Transmission lines and antennas will be treated in the chapters devoted to *Antennas;* the theory of operation of the various divisions of the transmitter proper will be discussed in the following pages.

The usual transmitter will contain the following general divisions: an oscillator, either crystal or self-controlled; one or more frequency multiplying stages; one or more radio frequency amplifying stages, and a system for either keying or modulating by voice the output of the final amplifier stage. However, a transmitter need not necessarily have all the stages mentioned above, and, in fact, may be merely an oscillator whose output is controlled by a telegraph key.

Oscillators

As was mentioned earlier, in the chapter devoted to the theory of vacuum tubes, the amplifying properties of a tube having three or more elements give it the ability to generate an alternating current of a frequency determined by the components associated with it. A vacuum tube operated in such a circuit is called an oscillator, and its function is essentially to convert a source of direct current into radio frequency alternating current of a predetermined frequency. Oscillators for controlling the frequency of conventional radio transmitters can be divided into two general classifications: self-controlled and crystal-controlled.

There are a great many types of self-controlled oscillators, each of which is best suited to a particular application. They again can further be subdivided into the classifications of: negative-grid oscillators, electron-orbit oscillators, and negative-resistance oscillators.

Negative-Grid Oscillators. A negative-grid oscillator is essentially a vacuum-tube amplifier with a sufficient portion of the output energy coupled back into the input circuit to sustain oscillation. They are called negative-grid oscillators because, in contrast to certain other oscillator circuits, the grid is biased a considerable amount negative with respect to the cathode. It is this classification of oscillator which finds most common application in low- and medium-frequency transmitter control circuits. The various common types of negative-grid oscillators are diagrammed in Figure 1.

The Hartley. Figure 1 (A) illustrates the oscillator circuit which finds the most general application at the present time; this circuit is commonly called the Hartley. The operation of this oscillator will be described as an index to the operation of all negative-grid oscillators; the only real difference between the various circuits is the manner in which energy for excitation is coupled from the plate to the grid circuit.

Operation of the Hartley Oscillator. When the plate voltage is applied to the plus and minus terminals of the Hartley oscillator shown at (A), the sudden flow of plate current accompanying the application of plate voltage will cause an electro-magnetic field to be set up in the vicinity of the coil. The building-up of this field will cause an instantaneous potential drop to appear from turn-to-turn along the coil. Due to the inductive coupling between the portion of the coil in which the plate current is flowing and the grid portion, a potential will be induced in the grid portion. Since the cathode tap is between the grid and plate ends of the coil, the induced grid voltage acts in such a manner as to increase further the plate current to the tube. This action will continue for a short period of time determined by the inductance and capacity of the tuned circuit, until the "flywheel"

effect of the tuned circuit causes this action to come to a maximum and then to reverse itself. The plate current then decreases, the magnetic field around the coil also decreasing, until a minimum is reached, when the action starts again in the original direction and at a greater amplitude than before. The amplitude of these oscillations, the frequency of which is determined by the coil-condenser circuit, will increase in a very short period of time to a limit determined by the plate voltage or the cathode emission of the oscillator tube.

The Colpitts. Figure 1 (B) shows a version of the Colpitts oscillator. It can be seen that this is essentially the same circuit as the Hartley except that a pair of capacitances in series are employed to determine the cathode tap, instead of actually using a tap on the tank coil. Also, the net capacity of these two condensers comprises the tank capacity of the tuned circuit.

The T.P.T.G. The tuned-plate tuned-grid or t.p.t.g. oscillator illustrated at (C) has a tank circuit in both the plate and grid circuits. The feedback of energy from the plate to the grid circuits is accomplished by the plate-to-grid interelectrode capacity within the tube.

For best operation of the Hartley and Colpitts oscillators, the voltage from grid to cathode, determined by the tap on the coil or the setting of the two condensers, should be from $\frac{1}{3}$ to $\frac{1}{5}$ that appearing between plate and cathode. In the t.p.t.g. oscillator, the grid circuit should be tuned to a frequency slightly lower than that of the plate circuit for best operation. The frequency of oscillation is determined primarily by the constants of the plate circuit, and therefore a broadly resonant or aperiodic coil may be substituted for the grid tank to form the *T.N.T.* oscillator shown at (D).

Electron-Coupled Oscillators. In any of the three oscillator circuits just described it is possible to take energy from the oscillator circuit by coupling an external load to the tank circuit. Since the tank circuit determines the frequency of oscillation of the tube, any variations in the conditions of the external circuit will be coupled back into the frequency determining portion of the oscillator. These variations will result in frequency instability.

Two oscillators, in which the frequency determining portion of the oscillator is coupled to the load circuit only by an electron stream, are illustrated in (E) and (F) of Figure 1. When it is considered that the screen of the tube acts as the plate to the oscillator circuit, the plate acting as a coupler to the load, then the similarity between the cathode-grid-screen circuit of these oscillators and the cathode-grid-plate circuits of the corresponding prototype can be seen.

The advantage of the electron-coupled oscillator over conventional types is in the greater stability with respect to load and voltage variations that can be obtained. Load variations have very little effect on the frequency of operation of the e.c.o., since the only coupling between the oscillating circuit and the load is through the electron stream flowing through the other elements to the plate. The plate is electrostatically shielded from the oscillating portion by the screen, which is at ground potential with respect to r.f.

The stability of the electron-coupled type of oscillator with respect to variations in supply voltages comes from an entirely different source. It is a peculiarity of such an oscillator that the frequency will shift in one direction with an increase in screen voltage while an increase in plate voltage will cause it to shift in the other direction. By a proper porportioning of the resistors that comprise the voltage divider supplying screen voltage, it is possible to make the frequency of the oscillator substantially independent of supply voltage variations; the tendency of a change in screen voltage to make the frequency shift in one direction is counterbalanced by the effect of the change in plate voltage to make the frequency shift in the other direction.

V. F. O. Transmitter Controls. During the last year or two there has been an increasing tendency to break away from the standard crystal oscillator as the only means of controlling the frequency of a transmitter because of the necessarily limited flexibility of such an oscillator. The new tendency has been toward the use of highly stabilized *variable-frequency oscillators* as transmitter controls in amateur equipment. These oscillators are nothing more than certain types of self-excited oscillators in which adequate precautions have been taken to insure that they shall be as stable as possible with respect to load and supply voltage variations.

Due to the better inherent stability of the electron-coupled type of oscillator, a number of the recent designs for *v.f.o.'s* (as the variable-frequency oscillators for transmitter frequency control are called) have used this type of oscillator. However, one disadvantage of the electron-coupled oscillator is that the cathode and heater are not at the same r.f. potential. This gives rise to difficulties due to heater-cathode leakage, heater-cathode capacity variation with changes in temperature, and coupling of stray r.f. energy from the heater into the cathode circuit.

As a consequence of this disadvantage of the electron-coupled oscillator, another group of

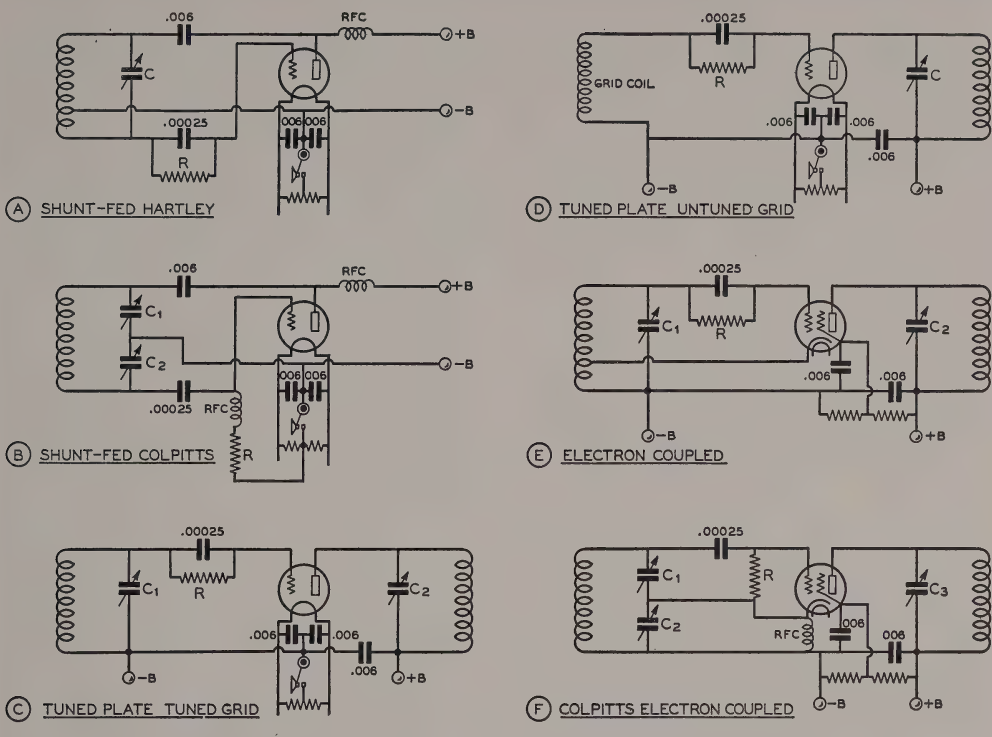


Figure 1.
COMMON TYPES OF SELF-EXCITED OSCILLATORS.

the recent designs for v.f.o.'s have used grounded-cathode oscillator circuits of the modified Hartley type. A v.f.o. of this design is shown in the chapter *Exciters and Low-Powered Transmitters.* Since the cathode of an oscillator of this type is at ground potential, it is impossible for r.f. energy from an external source to be coupled into the oscillating circuit from the heater circuit. However, the use of any type of oscillator as a transmitter control means that it must be carefully constructed, both from the electrical and from the mechanical standpoint.

Other Oscillator Circuits

Electron-Orbit Oscillators. Of the other oscillator circuits the negative-resistance and electron-orbit types are the most common of the self-excited class. Electron-orbit oscillators are used only for extremely high-frequency work (above 300 Mc.) and depend for their operation upon the fact that an electron takes a finite time to pass from one element to another inside a vacuum tube. The Gill-Morrell, Barkhausen-Kurtz, and Kozanowski oscillators are examples of this type and are described in the *Ultra-High Frequency Transmitters* chapter. Another special type of u.h.f. oscillator is the *magnetron,* which is also described in the *u.h.f.* chapter. This type employs a filament surrounded by a split plate to which are connected rods comprising a linear tank circuit. The tube is operated in a strong magnetic field; hence the name, magnetron.

Negative Resistance Oscillators. The other common type is the negative-resistance oscillator, which is used when unusually high frequency stability is desired, as in a frequency meter. The dynatron of a few years ago and the transitron of more recent fame are examples of oscillator circuits which make use of the negative resistance characteristic between different elements in a multi-grid tube. In the dynatron, the negative resistance is a consequence of secondary emission of electrons from the plate of the tube. By a proper proportioning of the electrode voltages, an increase in screen voltage will cause a decrease in screen current since the increased screen voltage will cause the screen to attract a larger number of the secondary electrons emitted by

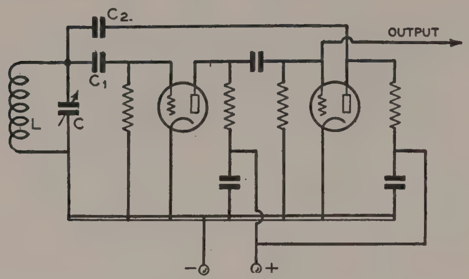


Figure 2.

THE FRANKLIN OSCILLATOR CIRCUIT.

In this oscillator, a separate phase-inverter tube is used to feed a portion of the output back into the input circuit in the proper phase to sustain oscillation.

the plate. Since the net screen current flowing from the screen supply will be decreased by an increase in screen voltage, it is said that the screen circuit presents a negative resistance.

If any type of tuned circuit, or even a resistance-capacitance circuit, is connected in series with the screen, the arrangement will oscillate—provided, of course, that the external circuit impedance is greater than the negative resistance. A negative resistance effect similar to the dynatron is obtained in the *transitron* circuit, which uses a pentode with the suppressor coupled to the screen. The negative resistance in this case is obtained from a combination of secondary emission and interelectrode coupling, and is considerably more stable than that obtained from uncontrolled secondary emission alone in the dynatron.

The Franklin Oscillator. A circuit which makes use of two cascaded tubes to obtain the negative-resistance effect is the Franklin oscillator, illustrated in Figure 2. The tubes may be either a pair of triodes, tetrodes, or pentodes, a dual triode, or a combination of a triode and a multi-grid tube. The chief advantages of this oscillator circuit are that only very loose coupling between the two tubes and the tank circuit, LC, is required, and that the frequency determining tank only has two terminals and one side of the circuit is grounded. Condensers C_1 and C_2 need be only 1 or 2 $\mu\mu$fd. for satisfactory operation of the oscillator; this means that tube capacity and input resistance variations will have only an extremely small effect on the oscillation frequency.

The Klystron Oscillator. A recent development in the field of ultra-high frequency electronics is the use of velocity-modulated electron beams. It is possible to velocity modulate the beam of electrons produced by an electron gun, and to couple electromagnetically an external load to this modulated stream of electrons. This is the principle of the "inductive output amplifier" of RCA. If, then, this velocity modulated stream of electrons is bounced back and forth inside a metal resonator, with a portion of the output being fed back to do the modulating of the electron beam, we have the klystron oscillator—a source of sizeable amounts of r.f. power in the micro-wave (less than 1 meter) region.

Crystal Controlled Oscillators

When it is desired to hold the frequency of a transmitter very closely to a certain definite value or to keep it within an assigned frequency tolerance, reliance is very commonly placed upon the *piezo-electric* properties of a plate cut from a natural crystal of quartz. Quartz crystals are very widely employed by amateurs and commercial services as frequency controls; hence, some of the important characteristics of piezo-electric minerals will be mentioned before entering into a discussion of the oscillators that make use of these characteristics for frequency control.

Quartz Crystals. Quartz and tourmaline are naturally occurring crystals having a structure such that when plates are cut in certain definite relationships to the crystallographic axes, these plates will show the piezo-electric effect—the plates will be deformed in the influence of an electric field, and, conversely, when such a plate is compressed or deformed in any way a potential difference will appear upon its opposite sides.

When such a quartz plate is placed in a circuit with a vacuum-tube amplifier having the output circuit coupled back into the input, and a tuned circuit in series with the plate of the amplifier tube, the circuit will self-oscillate at a frequency primarily determined by the frequency of mechanical resonance of the quartz plate. The frequency of mechanical resonance or frequency of oscillation of a quartz plate is dependent upon its physical dimensions and upon a constant determined by the crystallographic (or optical) cut of the plate. The stability of the frequency of oscillation of a crystal-controlled oscillator is dependent upon the Q of the quartz plate (determined by the optical cut, the accuracy of grinding, and the method of mounting) and upon the coefficient of temperature drift, which is determined primarily by the optical cut of the plate.

Crystal Cuts. The face of an X cut or Y cut crystal is made parallel to the Z axis in Figure 3. Special cut crystals, known as AT cut, V cut, LD2, HF2, etc., are cut with the

face of the crystal at an angle with respect to the Z axis, rather than being parallel to it. The purpose of the special cuts is to increase the power handling ability of the plates in some cases, but especially to reduce their temperature coefficient. AT, V, B5 and LD2 cut crystals have temperature coefficients approaching zero, and they should be used in radio transmitters in which accurate frequency control is essential. These crystals eliminate the need of a crystal oven for amateur work. A constant operating temperature is still required for many commercial applications, but the oven temperature need not be kept within as close limits as for an X or Y cut plate.

Spurious Peaks. Crystals that oscillate at more than one frequency are commonly known as crystals with multiple peaks. The dual vibrational tendency is more pronounced with Y cuts, but to a certain degree is exhibited by many X cuts. The use of a well designed, space-wound, low C tank coil in an oscillator will tend to discourage the crystal from oscillating at two frequencies, and in addition will increase the output. Experiments have shown that the frequency stability is not improved by large tank capacities, which only tend to augment the double frequency phenomenon.

Twin frequencies appear in several ways: sometimes the crystal will have two frequencies several hundred cycles apart, and will oscillate on both frequencies at the same time to produce an acoustically audible beat note. Other crystals will suddenly jump frequency as the tank tuning condenser is varied past a certain setting. Operation with the tank condenser adjusted near the point where the frequency shifts, is very unstable, the crystal sometimes going into oscillation on one frequency and sometimes on the other as the plate voltage is cut on and off. Still other crystals will jump frequency only when the temperature is varied over a certain range. And some plates will jump frequency with a change in either tank tuning or temperature.

Edge-of-Band Operation. When operating close to the edge of the band, it is advisable to make sure that the crystal will respond to but one frequency in the holder and oscillator in which it is functioning; any crystal with two peaks can jump frequency slightly without giving any indication of the change in the meter readings of the transmitter. If the transmitter frequency is such that operation takes place on the edge of the band at all times, under all conditions of room temperature, some form of temperature control will be required for the crystal unless it is of the zero drift type.

When working close to the edges of the 14 or 28 Mc. band, it is essential that the crystal temperature be kept at a fairly constant value; the frequency shift in kilocycles per degree increases in direct proportion to the operating frequency, regardless of whether the fundamental or harmonic is used. When a crystal shifts its frequency by 2 kilocycles, its second harmonic has shifted 4 kilocycles. Amateurs not operating on the edge of the band generally need not concern themselves about frequency drift due to temperature changes.

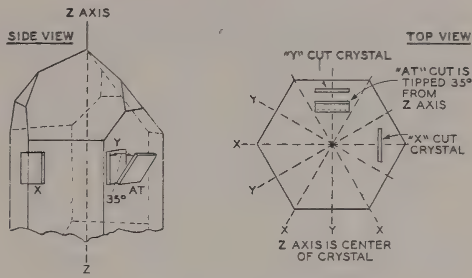


Figure 3.

SECTION THROUGH A RAW QUARTZ CRYSTAL.

Illustrating how crystals of various cut are removed from the raw quartz. Diamond saws usually are employed for cutting.

If a pentode or beam tetrode tube, having a plate potential of approximately 300 volts, is used for the crystal oscillator, the temperature of the crystal, regardless of cut, should not increase enough to cause any noticeable drift even at 14 Mc. When a crystal oscillator is keyed on 3.5 or 1.7 Mc., the frequency drift is not of any consequence, even with much higher values of plate input, because of the keying and of the fact that the drift is not multiplied as it would be with harmonic operation of a final amplifier.

The Crystal Holder. Crystal holders have a large effect on the frequency; for example, the frequency of an 80-meter crystal can vary as much as 3 kc. in different holders. In fact, crystals can be purchased in variable gap holders which enable the operator to vary the frequency by varying the air gap. From 20 to 50 kc. shift can be obtained at 14 Mc. with the newer types of variable gap crystals.

High-Frequency Crystals. Forty-meter crystals can be treated much the same as 80-meter crystals, provided they are purchased in a dust-proof holder from a reliable manufacturer. However, it is a good idea with 40-meter crystals to make sure that the crystal current is not excessive, as it will run higher in a given oscillator circuit than when a lower

frequency crystal is used in the same circuit at the same voltage. A low loss, low C tank circuit, and a pentode or beam type oscillator tube are desirable.

Third-Harmonic Crystals (14 and 28 Mc.). Twenty- and 10-meter crystals, especially the latter, require more care in regard to circuit details, components, and physical layout. These crystals are *not* of the zero drift type, as such crystals would be too thin to be of practical use. A special thick cut operated on a harmonic (almost always the third) is used to give the crystal sufficient mechanical ruggedness. Crystals of this cut have a drift of approximately 40-45 cycles/Mc./deg. C. This means that such crystals must be run at very low power levels, not only to avoid fracture, but to prevent excessive drift. However, their use permits considerable simplification of an u.h.f. transmitter.

A type 41 tube, running at 275 volts on the plate and 100 volts on the screen, makes a good oscillator tube for a 20-meter crystal. Bias should be obtained from a 500-ohm cathode resistor rather than from a grid leak. Very light loading, preferably with inductive coupling, is required. The tank coil should be low loss, preferably air-supported or wound on a ceramic form.

Medium high μ triodes with high transconductance and low input and output capacities make excellent 10-meter crystal oscillators. The types RK34, 6J5G, 7A4, and 955 are the most satisfactory oscillators, the 6J5G and 7A4 giving the greatest output besides being the least expensive.

Contrary to general practice with pentode crystal oscillators, the plate tank circuit should *not* be too low C; a moderate amount of tuning capacity should be used in a 10-meter triode crystal oscillator. The plate voltage on the oscillator tube should not be allowed to exceed 200 volts. About 2 watts output is obtainable from the 10-meter oscillator tank at this plate voltage. The tank coil can consist of 8 turns of no. 12 wire, air-wound to a $\frac{3}{4}$-inch diameter and spaced the diameter of the wire. Bias should be obtained from a 200-ohm cathode resistor (by-passed) and no grid leak. Connecting leads should be short, and components small physically.

Both 10- and 20-meter crystal oscillators should be followed, where practicable, by a tube of high power gain, such as the 807. This reduces the number of tubes required in a high power stationary u.h.f. transmitter.

A 10-meter crystal oscillator with a 6J5G, driving a 6V6G doubler using a 150,000-ohm grid leak, makes an excellent 5-meter mobile transmitter. The latter tube can be either plate or plate-and-screen modulated. The modulation is better, especially when doubling, if both plate and screen are modulated.

Crystal Oscillator Circuits

Crystal oscillators can be divided into three classifications: (1) low power circuits, which require several additional buffer stages to drive medium or high power final amplifiers; (2) high power crystal oscillators, which minimize the number of buffer stages in a transmitter; (3) harmonic crystal oscillators, which operate on more than one harmonically related band from one quartz crystal.

Low power crystal oscillators are often required in transmitter design where extremely stable frequency control is needed. The crystal oscillator tube is operated at low plate potential, such as 200 volts, with the result that oscillation is relatively weak. This means that there will be less heating effect in the quartz plate; the frequency drift, due to changes in temperature, is therefore minimized.

Mere operation of a quartz crystal oscillator tube at relatively low plate voltage does not necessarily mean a low degree of frequency drift; a type of crystal oscillator tube must be used which has high power sensitivity, high μ, and low feedback (interelectrode) capacity. The amount of feedback determines the value of r.f. current flowing through the quartz plate, and thus determines the amplitude of the physical vibration of the quartz plate. Any tube which requires only a very small amount of grid excitation voltage, and has low grid-to-plate capacity, can be used to supply relatively high-power output in a crystal oscillator without heating of the quartz plate.

High-power crystal oscillators are those which operate with as high a plate voltage as can be used with only moderate heating of the quartz crystal. Many transmitters, such as those used for amateur work, do not require as high a degree of frequency stability as do radiotelephone transmitters used for commercial services. The relatively high output from such crystal oscillators usually means the elimination of one or two buffer-amplifier stages. This simplifies the transmitter, and may result in more trouble-free operation. There are a great many types of tubes suitable for high-power crystal oscillators, some of which are also used in high-stability low-power crystal oscillators by merely reducing the electrode voltages.

The crystal oscillator circuit in Figure 4 is the standard oscillator circuit, and uses either a pentode or beam tetrode tube. It operates on one frequency only, and the plate circuit is

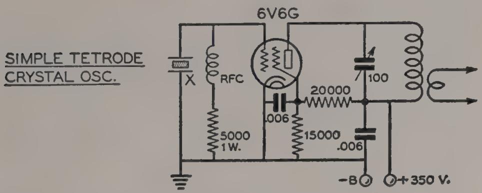


Figure 4.

TYPICAL CRYSTAL OSCILLATOR CIRCUIT.

This circuit has been found to be the most satisfactory for the frequency control of a multi-stage transmitter. A 6L6, 6F6, 42, or, for that matter, any pentode or power beam tetrode may be used with comparable success.

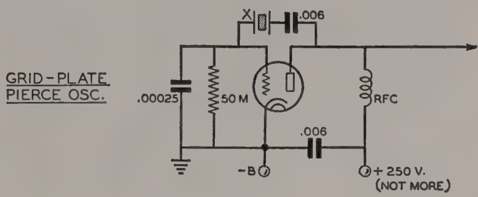


Figure 5.

TYPICAL PIERCE OSCILLATOR CIRCUIT.

No tank circuit is required with this type of crystal oscillator circuit. However, the crystal current is quite high for the amount of output voltage obtained.

tuned to a frequency somewhat higher than that of the quartz crystal.

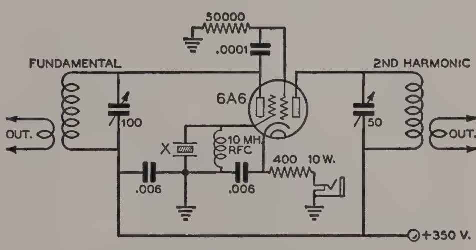
The actual power output of a crystal oscillator, such as shown in Figure 4, is from 1 to 15 watts, depending upon the values of plate and screen voltage. The use of *AT-cut* or low temperature coefficient quartz plates allows higher values of output to be obtained without exceeding the safe r.f. crystal current ratings or encountering frequency drift. *X-cut* and *Y-cut* crystals, especially the latter, must be operated with comparatively low crystal current because they not only will not stand as much r.f. crystal current, but also have a higher temperature coefficient.

Pierce Crystal Oscillator. One of the earliest crystal oscillator circuits recently enjoyed a revival in popularity. This is the Pierce oscillator, in which the crystal is connected directly from plate to grid of the oscillator tube, the crystal taking the place of the tuned tank circuit in an ultra-audion oscillator. Just as in the ultra-audion, the amount of feedback depends upon the grid to cathode capacity. Thus, it is only necessary to connect from grid to cathode a fixed condenser permitting the proper amount of feedback for the tube and frequency band used. The capacity is not at all critical, and ordinarily it is not necessary to change the capacity even when changing bands.

The chief advantage of the oscillator is that it requires no tuned circuits. The chief disadvantage is that the maximum obtainable output is low, due to the fact that not over 200-250 volts can be used safely. Also, it works well only with 160- and 80-meter crystals, though many 40-meter crystals will work satisfactorily if the constants are chosen for maximum performance on 40 meters.

The oscillator may be fed plate voltage either through an r.f. choke or a resistor of high enough resistance that it doesn't act as a low impedance path for the r.f. energy. A considerably higher power output can be obtained with an r.f. choke in the plate circuit, as compared to the use of a resistor in this position. However, since the plate voltage required on succeeding stages is invariably greater than that used on the Pierce crystal oscillator, the use of a resistor as the plate load is to be recommended. Figure 5 shows a popular version of the Pierce crystal oscillator.

Dual-Triode Oscillator-Doubler Circuits. The types 6N7, 6A6, and 53 twin-triode tubes are popular for circuits where one triode acts as a crystal oscillator which drives the other triode as a frequency doubler. One tube, therefore, serves a dual purpose, supplying approximately 5 watts output on either the fundamental frequency or the second harmonic of the quartz crystal. Two applications of the twin-triode tube in a crystal oscillator circuit are shown in Figures 6 and 7.



DUAL TRIODE OSCILLATOR DOUBLER

Figure 6.

TYPICAL TWIN-TRIODE OSCILLATOR-DOUBLER CIRCUIT.

Any dual triode of the 7F7, 6N7, 6F8G, 6A6, 53 class may be used in this simple circuit to obtain output on either the crystal frequency or its second harmonic.

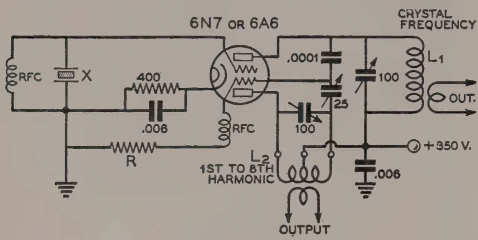


Figure 7.

REGENERATIVE DUAL-TRIODE OSCILLATOR.

By using one section of a dual triode as a crystal oscillator and the other section as a regenerative frequency multiplier, output on frequencies as high as the eighth harmonic of the crystal frequency may be obtained.

Figure 6 is a circuit which can be used with quartz crystals cut for 160-, 80-, 40- or 20-meter operation. The circuit shown in Figure 7 can be made regenerative in the frequency-multiplier section in order to use the second triode as a tripler or quadrupler. By reducing the capacity of the feedback condenser to a low enough value, the second triode can be neutralized for use as a buffer stage. A suitable condenser for this purpose is a small mica-insulated trimmer condenser having a capacity range of from 3-to-30$\mu\mu$fds.

The resistor R shown in Figure 7 should be from 30,000 to 50,000 ohms in value. The r.f. choke shown in series with this resistor frequently can be omitted.

Harmonic Oscillator Circuits. Harmonic oscillator circuits can be generally defined as those crystal oscillator arrangements which use a single tube and which allow power output to be obtained on harmonics of the crystal frequency. While these oscillator circuits have the advantage that one or more tubes are eliminated from the lineup, and sometimes that a tuned circuit is eliminated, they all have the disadvantage that they are difficult to adjust properly, and they all have a tendency toward excessive crystal current when improperly tuned up. Five of the best known and most satisfactory of these oscillator circuits have been grouped together in Figure 8.

The Tritet Crystal Oscillator. Any of the common pentode, tetrode, or screen-grid tubes may be used in the tritet crystal oscillator as shown in Figure 8A. There are really two active circuits in this oscillator arrangement: the grid-cathode-screen circuit, which acts as a triode crystal oscillator, and the cathode-grid-plate circuit, which acts as an r.f. amplifier or frequency multiplier, with its output circuit shielded from the oscillator portion. The tetrode or pentode plate circuit is *electron coupled* to the oscillator circuit. The plate circuit is generally tuned to the second harmonic, and outputs of from 5 to 15 watts can be obtained without damage to the quartz crystal. This circuit is an improvement over the older forms of tritet in which a grid leak was used in place of the grid r.f. choke, and in which no cathode resistor and by-pass condenser were included. The improved circuit (Figure 8A) decreases the crystal current as much as 50 per cent, and thereby protects the crystal against fracture. The cathode circuit is high C, and is tuned to a frequency which is 40 to 50 per cent higher than that of the crystal. If an 802 or 807 is substituted for the 6L6 tube, the plate circuit can be tuned to the fundamental frequency of the crystal without making it necessary to short-circuit the cathode tuned circuit. A further reduction in r.f. crystal current may be obtained by connecting a 140-$\mu\mu$fd. variable condenser between the bottom of the crystal and the top of the cathode tank coil L_2. This condenser should be set to the smallest value of capacity which will permit steady oscillation and full output.

Regenerative Oscillator Circuits. Figures 8B and 8C show two versions of a regenerative crystal oscillator circuit, which requires only one tank circuit, and which is capable of giving power output on harmonics of the crystal frequency. Figure 8B shows the circuit for use with a triode tube such as the 76, 6C5, 6J5, and 7A4, given in the order of their efficacy. Figure 8C shows the same circuit adapted for use with a pentode or beam tetrode, such as the 42, 6F6, 6V6, 6L6, or 7C5.

Triodes such as the 7A4, 6J5-GT, and the 76 will deliver as much as 2 or 3 watts with an r.f. crystal current of between 10 and 60 ma. for crystals from 160 to 10 meters. The triode circuit is excellent to drive a 6L6G buffer-doubler, and the screen supply voltage for the 6L6G tube may be applied to the 76 plate circuit. This type of circuit is the only one which works with all crystals, 10, 20, 40, 80 and 160 meters, whether they are extremely active, such as a good X cut, or relatively inactive, such as most high-frequency crystals. The triode will furnish from 1 to 2 watts at twice crystal frequency when used with 160-, 80-, or 40-meter crystals by tuning the plate circuit to the second harmonic.

In Figure 8B the cathode condenser, C_1, usually is left at some setting of from 40 to 50 $\mu\mu$fd. for 40-, 80-, and 160-meter crystals.

A 6F6 or 42 works very well in the Figure 8C circuit with a C_1 value of .0001 μfd. if heavily loaded. Eight to 12 watts output can be obtained easily from 160 to 20 meters, and

about 5 watts on 10 meters. A 6L6G tube requires a higher value of C_1, about .0004 μfd. unless heavily loaded.

Reinartz Crystal Oscillator. The Reinartz 802 crystal oscillator has a fix-tuned cathode circuit which is resonated to approximately *one-half* the crystal frequency. For example, with an 80-meter crystal the cathode circuit is tuned to 160 meters, the plate circuit to 80 meters. Either an 802 or a 6F6 tube can be used in a Reinartz crystal oscillator circuit. The output will be from 5 to 25 watts, depending upon the values of plate and screen voltages. The 6F6 is used as a high-μ triode in this same type of circuit, whereas the 802 is used as a pentode oscillator with additional control grid-to-plate capacity feedback. The circuit is shown in Figure 8D.

The crystal r.f. current is quite low in this circuit, in comparison with the output power which can be obtained. The cathode circuit is tuned to half the frequency of the crystal, and the reactive effect produces regeneration at the harmonic frequency. This increases the operating efficiency of the tube without danger of uncontrollable oscillation at frequencies other than that of the crystal.

Improved Harmonic Oscillator. Figure 8E shows an improved version of a harmonic oscillator arrangement which has been suggested by Jones. It is quite similar to previous arrangements in regard to the general hookup, but in this arrangement the screen is by-passed back to the cathode (which is hot to r.f.) rather than to ground. This is said to increase the stability of the oscillator, and to increase the efficiency of the arrangement when operating on harmonics of the crystal frequency.

Push-Pull Crystal Oscillators. Figure 9 shows a simple crystal oscillator arrangement which makes use of one of the common dual-triode tubes as a push-pull oscillator. The type 6A6, 6N7, and 7F7 dual triodes make good push-pull crystal oscillators.

Outputs of from 5 to 10 watts can be obtained from this circuit without exceeding the

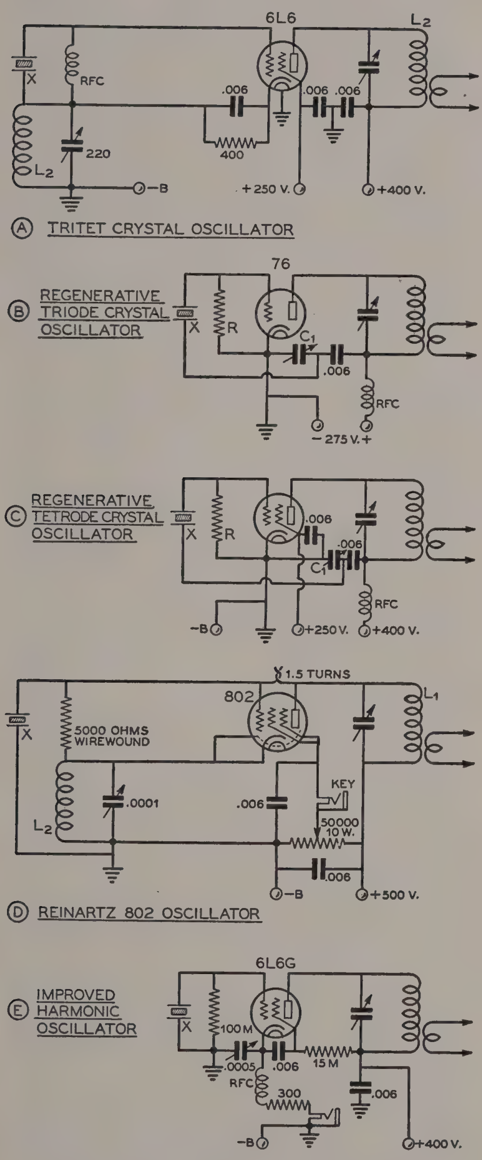


Figure 8.

REGENERATIVE OSCILLATOR CIRCUITS.

Full details of the operation of these oscillator circuits and a comparison between them is given in the text.

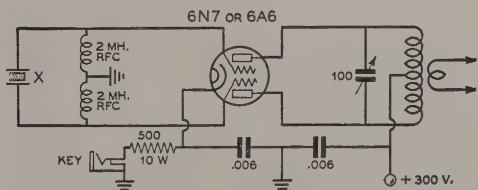


Figure 9.

PUSH-PULL CRYSTAL OSCILLATOR.

Any of the dual-triode tubes make a quite satisfactory push-pull crystal oscillator for feeding a push-pull r.f. amplifier, a push-push doubler, or merely to obtain somewhat greater output than from a single-ended oscillator.

ratings of the usual X-cut crystals. The crystal current for a push-pull oscillator is but little higher than for a single triode of the same type, and greater power output can be obtained.

Some push-pull oscillators will not oscillate on 160 meters, the feedback being insufficient in the push-pull connection to sustain oscillation under load.

Tuning the Crystal Oscillator

In nearly every practical transmitter circuit there will be some means for determining proper tuning of the crystal oscillator stage. Perhaps the most satisfactory of these tuning indicators is the grid milliammeter of the following stage. Maximum meter reading indicates maximum output from the crystal oscillator. Other indicators are: (1) A small neon bulb held near the plate end of the oscillator tuned circuit; maximum glow of the bulb indicates maximum oscillator output. (2) A flashlight bulb or a pilot light bulb, connected in series with a turn of wire fastened to a long piece of wood dowel (to protect the operator), can be coupled to the oscillator coil for indicating r.f. output. Maximum brilliancy of the lamp denotes maximum power output from the oscillator.

Oscillator-Doubler Circuits. The type 6N7 or 6A6 oscillator-doubler circuit is adjusted by tuning the oscillator section for maximum output, and the doubler section for greatest dip in cathode or plate current. The crystal plate section should generally be tuned until the circuit approaches the point where oscillation is about to cease; this is towards the higher-capacity setting of the oscillator plate tuning condenser, and operation in this manner provides most output in proportion to r.f. crystal current and frequency drift.

Harmonic Oscillators. Harmonic crystal oscillators are always tuned for maximum output and minimum plate, or cathode current. The regeneration or feedback condenser is adjusted or chosen to provide a good plate current dip when the plate circuit is tuned to the second harmonic of the crystal oscillator. Too much regeneration will cause the tube to oscillate for all settings of the plate tank condenser, without any sharp dip at the harmonic frequency of the crystal. Insufficient regeneration will result in low second harmonic output.

A plate potential of 400 volts is generally considered a safe upper limit for a type 6L6 oscillator tube. The screen-grid voltage affects the degree of regeneration and harmonic output; this voltage should generally range between 250 and 275 volts. The cathode current will run between 50 and 60 milliamperes for fundamental frequency operation, and 60 to 75 milliamperes for harmonic operation, at these plate and screen voltages. The crystal r.f. current normally runs between 25 and 75 milliamperes in this type of oscillator, depending on the frequency and plate voltage used.

Radio-Frequency Amplifiers

Since the output of the oscillator stage in a transmitter (whether it be self-controlled or crystal controlled) must be kept down to a fairly low level to maintain stability and to maintain a factor of safety from fracture of the crystal when one is used, the low power output of the oscillator is brought up to the desired power level by means of radio-frequency amplifiers. The two classes of amplifiers that find widest application in amateur transmitters are the class B and class C types.

The Class B Amplifier. Class B amplifiers are used in a radio-telegraph transmitter when maximum power gain is desired in a particular stage. A class B amplifier operates with cutoff bias and a comparatively small amount of excitation. Power gains of 20 to 200 or so are obtainable in a well-designed class B amplifier. The plate efficiency of a class B c.w. amplifier will run around 65 per cent.

The Class B Linear. Another type of class B amplifier is the class B linear stage as employed in radiophone work. This type of amplifier is used to increase the level of a modulated carrier wave, and depends for its operation upon the linear relation between excitation voltage and output voltage. Or, to state the fact in another manner, the power output of a class B linear stage varies linearly with the square of the excitation voltage. The class B linear amplifier is operated with cutoff bias and a small value of excitation, the actual value of exciting power being such that the power output under carrier conditions is one fourth of the peak power capabilities of the stage. Class B linears are very widely employed in broadcast and commercial installations, but are comparatively uncommon in amateur application, since tubes with high plate dissipation are required for moderate output. Another reason for their unpopularity among amateurs is that the power limitation upon amateurs is placed upon power *input* to the final stage, and not upon power *output*. The approximately 33 per cent efficiency of the class B linear makes the power capability of a transmitter with a linear amplifier in the final stage less than half that of a high-level modulated transmitter, whose maximum efficiency may be as high as 75 or 80 per cent. This assumes that the maximum legal input of 1 kilowatt is being employed in each case.

The Class C Amplifier. Class C amplifiers are very widely employed in all types of transmitters. A good power gain may be obtained (values of gain from 3 to 20 are common) and the plate circuit efficiency may be, under certain conditions, as high as 85 per cent. Class C amplifiers operate with considerably more than cutoff bias and ordinarily with a rather large amount of excitation as compared to a class B amplifier. The bias for a normal class C amplifier is such that plate current on the stage flows for approximately 120° of the 360° excitation cycle. Class C amplifiers are used in transmitters where a fairly large amount of excitation power is available, where good plate circuit efficiency is desired, and when the stage is to be plate modulated.

Class C Plate Modulation. The characteristic of a class C amplifier which makes it linear with respect to changes in plate voltage is that which allows such an amplifier to be *plate modulated* for radiotelephony. Through the use of higher bias than is required for a c.w. class C amplifier and greater excitation, the linearity of such an amplifier may be extended from zero plate voltage to twice the normal value. The output power of a class C amplifier, adjusted for plate modulation, varies with the square of the plate voltage. Since this is the same condition that would take place if a resistor equal to the voltage on the amplifier, divided by its plate current, were substituted for the amplifier, it is said the stage presents a resistive load to the modulator.

Class C Grid Modulation. If the grid current to a class C amplifier is reduced to a low value, and the plate loading is increased to the point where the plate dissipation approaches the rated value, such an amplifier may be grid modulated for radiotelephony. If the plate voltage is raised to quite a high value and the stage is adjusted carefully, efficiencies as high as 42 to 45 per cent with good modulation capability and comparatively low distortion may be obtained. This type of operation is termed class C grid modulation and is coming into increasing favor among amateur radiotelephone operators.

Grid Excitation. A sufficient amount of grid excitation must be available for class B or class C service. The excitation for a plate-modulated class C stage must be sufficient to drive a normal value of d.c. grid current through a grid bias supply of about $2\frac{1}{2}$ times cutoff. The bias voltage preferably should be obtained from a combination of grid leak and fixed C-bias supply. Cutoff bias can be calculated by dividing the amplification factor of the tube into the d.c. plate voltage. This is the value normally used for class B amplifiers (fixed bias, no grid leak). Class C amplifiers use from $1\frac{1}{2}$ to 5 times this value, depending upon the available grid drive, or excitation, and the desired plate efficiency. Less grid excitation is needed for c.w. operation, and the values of fixed bias (if greater than cutoff) may be reduced, or the value of the grid leak resistor can be lowered until normal d.c. grid current flows. This value should be between 75 per cent and 100 per cent of the value listed under tube characteristics.

The values of grid excitation listed for each type of tube may be reduced by as much as 50 per cent if only moderate power output and plate efficiency are desired. When consulting the tube tables, it is well to remember that the power lost in the tuned circuits must be taken into consideration when calculating the available grid drive. At very high frequencies, the r.f. circuit losses may even exceed the power required for grid drive unless low loss tank circuits are used.

Readjustments in the tuning of the oscillator, buffer, or doubler circuits, will often result in greater grid drive to the final amplifier. The actual grid driving power is proportional to the d.c. voltage developed across the grid leak (or bias supply), multiplied by the d.c. grid current.

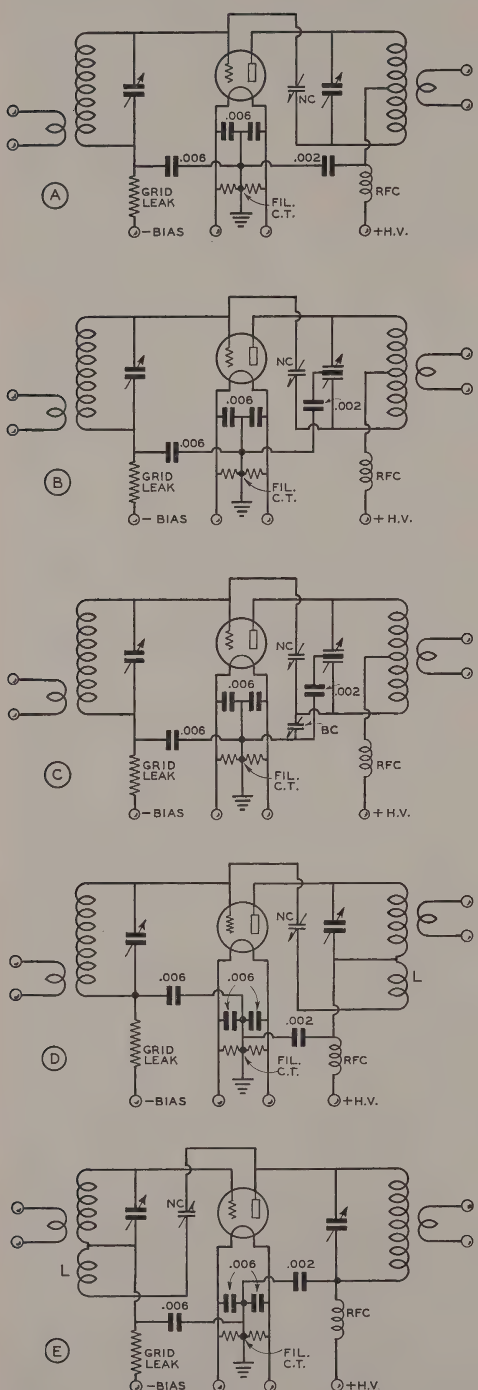
Link coupling between stages, particularly to the final amplifier grid circuit, normally will provide more grid drive than can be obtained from other coupling systems. The number of turns in the coupling link, and the location of the turns on the coil can be varied with respect to the tuned circuits to obtain the greatest grid drive for allowable values of buffer or doubler plate current. Slight readjustments sometimes can be made after plate voltage has been applied to the driver tube.

Excessive grid current will damage the tubes by overheating the grid structure; beyond a certain point of grid drive, no increase in power output can be obtained for a given plate voltage.

Neutralization of R. F. Amplifiers

The plate-to-grid feedback capacity of triodes makes it necessary that they be neutralized for operation as r.f. amplifiers at frequencies above about 500 kc. Those screen-grid tubes, pentodes, and beam tetrodes which have a plate-to-grid capacity of a small fraction of 1 $\mu\mu$fd. may ordinarily be operated as an amplifier without neutralization.

Neutralizing Circuits. The object of a neutralization circuit for an r.f. amplifier is, of course, to cancel or "neutralize" the capacitive feedback of energy from plate to grid. There are two general methods by which this energy



feedback may be eliminated: the first, and the most common method, is through the use of a capacity bridge, and the second method is through the use of a parallel reactance of equal and opposite polarity to the grid-to-plate capacity, to nullify the effect of this capacity.

Until recently, the capacity-bridge method of neutralization was divided into two systems: grid neutralization and plate neutralization. It has always been known that the use of grid neutralization caused an amplifier to be either regenerative or degenerative, but it was not until quite recently that Doherty showed the reason for the unsatisfactory performance of capacity bridge grid neutralization. Hence, only plate neutralization (the capacity bridge system), coil neutralization (the opposite reactance system), and Hazeltine neutralization will be considered as satisfactory methods for neutralizing a single-ended r.f. amplifier stage.

Tapped-Coil Plate Neutralization. As was mentioned under *Neutralizing Circuits*, there are two general types of neutralizing circuits for a single-ended amplifier: the bridge and opposite-reactance methods. The following paragraphs will describe first the variations upon the bridge method. Figure 10A shows a circuit for the neutralization of a single-ended triode r.f. amplifier by means of a tapped coil in the plate circuit. This circuit is satisfactory for frequencies below about 7 Mc. with ordinary tubes, but a considerable amount of regeneration will be found when this

Figure 10.

SINGLE-ENDED AMPLIFIER NEUTRALIZING CIRCUITS.

(A) shows a neutralizing circuit employing a split-coil plate tank which is suitable under ordinary conditions for operation at frequencies as high as 7 Mc. (B) shows conventional split-stator plate neutralization. (C) shows split-stator plate neutralization with the addition of a balancing condenser BC which compensates for the plate-to-ground capacity of the output tube and thus keeps the circuit balanced to ground. (D) shows the plate-neutralized Hazeltine circuit. The coil L is wound adjacent to the plate coil, but is separate from it and can be wound of smaller size wire. The larger the coil, the smaller will be the neutralizing condenser NC. (E) shows the grid-neutralized Hazeltine circuit. The same conditions apply to L and NC as were discussed under (D) above. This circuit has the advantage, however, that the high-power plate tank is isolated from the neutralizing circuit; the neutralizing coil is included on the relatively low-powered grid tank circuit.

circuit is used on frequencies above 7 Mc. Some regeneration can be tolerated in an amplifier for c.w. use, but for phone operation, either of the split-stator circuits described in the next two paragraphs should be used.

Split-Stator Plate Neutralization. Figure 10B shows the neutralization circuit which is most widely used in single-ended r.f. stages. The use of a split-stator plate condenser makes the electrical balance of the circuit substantially independent of the mutual coupling within the coil and also makes the balance independent of the place where the coil is tapped. With conventional tubes this circuit will allow one neutralization adjustment to be made on, say, 14 Mc., and this adjustment will hold for all lower frequency bands.

Capacity-Balanced Split-Stator Plate Neutralization. Figure 10C shows an alternative circuit for split-stator neutralization of a single-ended amplifier stage which, with low-capacity tubes, can be made to remain in adjustment on all bands from 56 Mc. on down in frequency. The additional balancing condenser BC serves merely as an adjustment to keep the capacity-to-ground exactly the same from each side of the balanced plate tank circuit. This condenser can be either a small adjustable one of the type commonly used for neutralization, or the relative capacity to ground of the two sides of the circuit can be proportioned so that there is a balance. In determining the balance of the circuit, it must be remembered that the plate-to-filament capacity of the power amplifier tube is the main item to cause the unbalance. If the other capacities of the circuit are perfectly balanced with respect to ground, the capacity of the condenser BC should be approximately equal to the plate-to-ground capacity of the tube being neutralized. However, it is often just as convenient to unbalance the circuit capacities to ground until the additional capacity on the neutralizing side of the circuit is about equal to that on the plate side. At the point where the plate-to-ground capacity is exactly balanced, the amplifier will neutralize perfectly (at least as nearly perfect as a push-pull amplifier) and will stay neutralized on all bands for which the amplifier tubes are satisfactory.

Hazeltine Neutralization. An alternative system of neutralization, wherein the neutralizing circuit is inductively coupled to one of the tank coils, is shown in Figures 10D and 10E. Figure 10D shows the plate neutralized Hazeltine circuit, while 10E shows the grid neutralized arrangement. In either case, it will be noticed that there is no tank current flowing through the neutralizing coil L.

In this circuit arrangement, the size of the neutralizing condenser NC is determined by the coefficient of coupling between the tank coil and L, and upon their relative inductances. It is possible, by proper proportioning of the neutralizing coil L on each band, to make one setting of NC correct for all bands.

Push-Pull Neutralization. Two tubes can be connected for *push-pull* operation so as to obtain twice as much output as that of a single tube. A push-pull amplifier, such as that shown in Figure 11A, also has an advantage in that the circuit can more easily be balanced than a single-tube r.f. amplifier. The various interelectrode capacities and the neu-

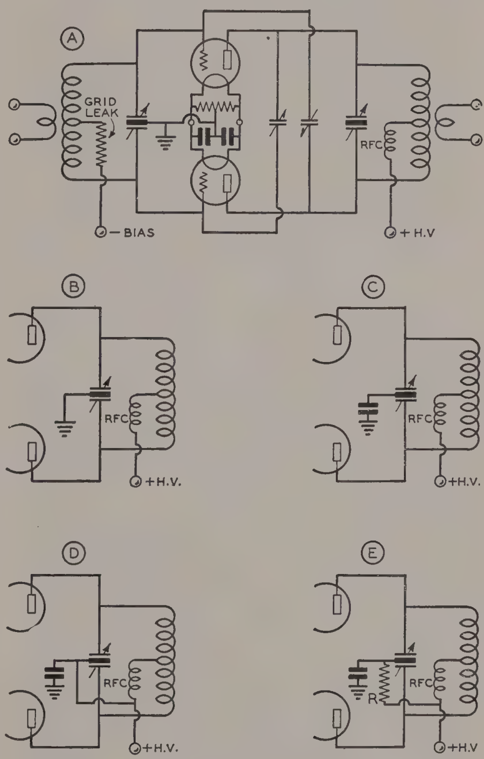


Figure 11.

PUSH-PULL AMPLIFIER NEUTRALIZATION.

(A) shows the basic circuit for a neutralized push-pull r.f. amplifier. In this circuit the nodal point for the stage is determined by the grounded rotor on the grid tuning condenser, and the rotor of the plate tank condenser is allowed to float. (B), (C), (D), and (E) show alternative arrangements for returning the rotor of the plate tank condenser to ground when this grounding is deemed necessary. Discussion of the various circuits is given in the text.

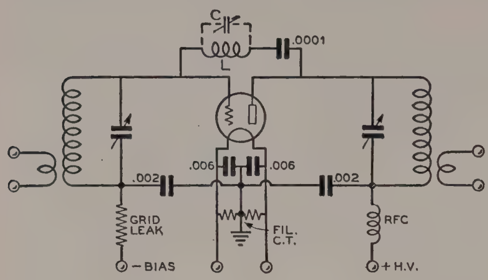


Figure 12.

SHUNT OR "COIL" NEUTRALIZATION.

This neutralization circuit makes use of a coil connected from grid to plate (with a blocking condenser in series with it) which resonates with the grid-to-plate capacity to the operating frequency. The impedance from plate to grid is thus made very high, feedback is stopped, and the amplifier is neutralized for this frequency of operation. When the frequency of operation is changed, the trimmer condenser C is adjusted to change the resonant frequency of this circuit to the new operation frequency.

tralizing condensers are connected in such a manner that those on one side of the tuned circuits are exactly equal to those on the opposite side. For this reason, push-pull r.f. amplifiers can be more easily neutralized in very-high-frequency transmitters; also, they usually remain in perfect neutralization when tuning the amplifier to different bands.

The circuit shown in Figure 11A is perhaps the most commonly used arrangement for a push-pull r.f. amplifier stage. The rotor of the grid condenser is grounded, and the rotor of the plate tank condenser is allowed to float. Under certain conditions, the circuit of 11B may be used (when the plate tank condenser has a much larger voltage rating than the maximum possible peak output of the power tubes) with the rotor of the grid condenser grounded or not, as desired. It is also possible to use a single-section grid condenser with a tapped coil (un-bypassed) for low-frequency operation with this circuit arrangement.

Figure 11C shows an alternative arrangement for the return of the rotor of the plate tank condenser which is best for use with a c.w. amplifier stage. The by-pass condenser from the rotor to ground can be any capacity from .01 μfd. down to .0005 μfd. and even down to .0001 μfd. for an u.h.f. amplifier. For phone use, it is best to have some sort of a coupling arrangement to make the rotor of the tuning condenser follow plate voltage fluctuations. As long as the rotor of the tuning condenser is at the same d.c. potential as the stators, there will be a much reduced chance of breakdown on modulation peaks.

Figures 11D and 11E show two arrangements which tend to keep the rotor of the condenser as nearly as possible at the same d.c. potential as the stators. In Figure 11D the rotor of the condenser, and the ungrounded side of the by-pass condenser, is merely connected to the plate supply side of the r.f. choke. This is an excellent arrangement for use with moderate plate voltages but has the disadvantage that considerable stress is placed on the mica by-pass condenser, and, should this condenser break down, the plate supply would be shorted. Figure 11E shows an alternative arrangement which has the advantage that, should the mica by-pass condenser short out, only the resistor R will be destroyed. For a mica by-pass capacity of .001 μfd. and a maximum 100 per cent modulation frequency of 3000 cycles, a 25,000-ohm resistor will be satisfactory for R.

Shunt Neutralization. The feedback of energy from grid to plate that would cause oscillation or serious regeneration in an unneutralized r.f. amplifier is a result of the grid-to-plate capacity of the amplifier tube. A neutralization circuit is merely an electrical arrangement for nullifying the effect of this capacity. All the previous neutralization circuits have made use of a bridge circuit for balancing out the grid-to-plate energy feedback by feeding back an amount of energy equal and opposite in phase.

Another method of eliminating the feedback effect of this capacity, and hence of neutralizing the amplifier stage, is shown in Figure 12. The grid-to-plate capacity in the triode amplifier tube acts as a capacitive reactance, coupling energy back from the plate to the grid circuit. If we parallel this capacity with an inductance having the same value of reactance (but having the opposite sign, of course) at the frequency upon which the amplifier is operating, the reactance of one will cancel the reactance of the other and we will have a high-impedance tuned circuit from grid to plate on the triode tube.

This neutralization circuit works very beautifully and can be used on ultra-high frequencies where other neutralization circuits are unsatisfactory. This is true because the lead length in the neutralization circuit is practically negligible. The circuit can also be used with push-pull r.f. amplifiers; in this case each tube is neutralized separately although both neutralizing condensers are set to the same capacity.

The big advantage of this arrangement is

that it allows the use of single-ended tank circuits with a single-ended amplifier.

However, the circuit has one serious disadvantage for amateur work in which the frequency of operation is changed frequently: the neutralization holds for one frequency—that frequency where the grid-to-plate capacity is resonant with the external neutralization coil. But by the use of plug-in coils and the trimmer condenser C in parallel with the grid-to-plate capacity, it is possible to shift the band of operation and to trim to any frequency within the band. This trimmer condenser, if used, must be insulated for somewhat more voltage than the tank condenser. The .0001-μfd. condenser in series with the neutralizing circuit is merely a blocking condenser to isolate the plate voltage from the grid circuit. The coil L will have to have a very large number of turns for the band in operation in order to be resonant with the usually rather small grid-to-plate capacity. But since, in all ordinary cases with tubes operating on frequencies for which they were designed, the L/C ratio of the tuned circuit will be very high, the coil can use comparatively small wire, although it must be wound on air or very low-loss dielectric, and must be insulated for the sum of the plate r.f. voltage and the grid r.f. voltage.

Figure 13 shows an alternative arrangement for the neutralizing circuit in which the variable trimmer condenser is in series with the neutralizing coil instead of in parallel with it. This system also allows the stage to be trimmed to neutralization on any frequency in the band of operation. This condenser can have a capacity of 5 to 25 $\mu\mu$fd. and need not have nearly as much voltage insulation as the trimmer condenser shown in Figure 12. A plate spacing of .070 of an inch will be ample for any plate voltage ordinarily used by the amateur.

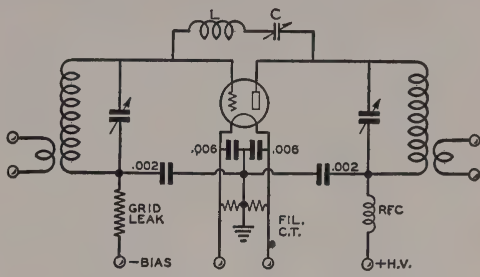


Figure 13.

ALTERNATIVE SHUNT NEUTRALIZATION CIRCUIT.

In this circuit the trimmer condenser for varying the frequency of resonance of the circuit is placed in series with the neutralizing coil, thus replacing the blocking condenser and reducing the necessary voltage rating for the trimmer condenser, although increasing the capacity required.

Neutralizing Procedure

The r.f. amplifier is neutralized to prevent self-oscillation or regeneration. A neon bulb, a flashlight lamp and a loop of wire, or an r.f. galvanometer can be used as a *null indicator* for neutralizing low-power stages. *The plate voltage lead is disconnected from the r.f. amplifier stage while it is being neutralized.* Normal grid drive then is applied to the r.f. stage, the neutralizing indicator is coupled to the plate coil, and the plate tuning condenser is tuned to resonance. The neutralizing condenser (or condensers) then can be adjusted until *minimum* r.f. is indicated for resonant settings of both grid and plate tuning condensers. Both neutralizing condensers are adjusted simultaneously and to approximately the same value of capacity when a push-pull stage is being neutralized.

A final check for neutralization should be made with a d.c. milliammeter connected in the grid leak or grid-bias circuit. There will be no movement of the meter reading as the plate circuit is tuned through resonance (without plate voltage being applied) when the stage is completely neutralized. The milliammeter check is more accurate than any other means for indicating complete neutralization and it also is suitable for neutralizing the stages of a high-power transmitter.

Push-pull circuits usually can be more completely neutralized than single-ended circuits when operating at very high frequencies. In the intermediate range of from 3 to 15 Mc., single-ended circuits will give satisfactory results.

Neutralizing Problems. When a stage cannot be completely neutralized, the difficulty can be traced to one or more of the following causes: (1) The filament leads may not be by-passed to the common ground bus connection of that particular stage. (2) The ground lead from the rotor connection of the split-stator tuning condenser to filament may be open or too long. (3) The neutralizing condensers may be in a field of excessive r.f. from one of the tuning coils. (4) Electromagnetic coupling may exist between grid and plate coils, or between plate and receding buffer or oscillator circuits. (5) Insufficient shielding or spacing between stages, or between grid and plate circuits in compact transmitters, may prevent neutralization or give false indications of neutralizing adjustments. (6) If shielding

is placed too close to plate circuit coils, neutralization will not be secured because of induced currents in the shields. (7) Parasitic oscillations may take place when plate voltage is applied. The cure for the latter is mainly a matter of cut and try—rearrange the parts, change the length of grid or plate or neutralizing leads, insert an ultra-high-frequency r.f. choke in the grid lead or leads, or eliminate the grid r.f. chokes which may be the cause of a low-frequency parasitic (in conjunction with plate r.f. chokes). See *Parasitic Oscillation in R.F. Amplifiers* later in this chapter.

Plate Circuit Tuning. When the amplifier is completely neutralized, reduced plate voltage should be applied before any load is coupled to the amplifier. This reduction in plate voltage should be at least 50 per cent of normal value because the plate current will rise to excessive values when the plate tuning condenser is not adjusted to the point of resonance. The latter is indicated by the greatest dip in reading of the d.c. plate current milliammeter; the r.f. voltage across the plate circuit is greatest at this point. With no load, the r.f. voltage may be several times as high as when operating under conditions of full load; this may result in condenser flashover if normal d.c. voltage is applied. The no-load plate current at resonance should dip to 10 per cent or 20 per cent of normal value. If the plate circuit losses are excessive, or if *parasitic oscillations* are taking place, the no-load plate current will be higher.

Loading. The load (antenna or succeeding r.f. stage) then can be coupled to the amplifier under test. The coupling can be increased until the plate current at resonance (greatest dip in plate current meter reading) approaches the normal values for which the tube is rated. The value at reduced plate voltage should be proportionately less in order to prevent excessive plate current load when normal plate voltage is applied. Full plate voltage should not be applied to an amplifier unless the r.f. load also is connected; otherwise the condensers will arc or flash over, thereby causing an abnormally high plate current which may damage the tube. The tuned circuit impedance is lowered when the amplifier is loaded, as are the r.f. voltages across the plate and neutralizing condensers.

Grid Excitation. Excessive grid excitation is just as injurious to a vacuum tube as abnormal plate current or low filament voltage. Too much grid driving power will overheat the grid wires in the tube, and will cause a release of gas in certain types of tubes. An excess of grid drive will not appreciably increase the power output and can increase the efficiency only slightly after a certain point is reached. The grid current in the tube should not exceed the values listed in the *Tube Tables,* and care also should be exercised to have the bias voltage low enough to prevent flashover in the stem of the vacuum tube.

Grid excitation usually refers to the actual r.f. power input to the grid circuit of the vacuum tube, part of which is used to drive the tube, and part of which is lost in the C-bias supply. There is no way to avoid wasting a portion of the excitation power in the bias supply.

Frequency Multipliers

Quartz crystals are not ordinarily used for direct control of the output of high-frequency transmitters. *Frequency multipliers* are usually needed to multiply the frequency to the desired value. These multipliers operate on exact multiples of the crystal frequency; a 3.6-Mc. crystal oscillator can be made to control the output of the transmitter on 7.2 or 14.4 Mc., or even on 28.8 Mc., by means of one or more frequency multipliers. When used at twice frequency, as they most usually are, they are often termed *frequency doublers.* A simple doubler circuit is shown in Figure 14. It consists of a vacuum tube with its plate circuit tuned to *twice* the frequency of the grid driving circuit. This doubler can be excited from a crystal oscillator, or connected to another doubler or buffer amplifier stage.

Doubling is best accomplished by operating the tube with extremely high grid bias in order to make the output plate current rich in harmonics. The grid circuit is driven approximately to the normal value of d.c. grid current through the r.f. choke and grid leak resistor, shown in Figure 14. The resistance value generally is from two to five times as high as that used with the same tube for simple amplification. Consequently, the grid bias is several times as high for the same value of grid current.

Neutralization is seldom absolutely necessary in a doubler circuit, since the plate is tuned to twice the frequency of the grid circuit. The feedback from the doubler plate circuit to the grid circuit is at *twice* the frequency of the grid driving circuit to which the coupling condenser (Figure 14) is connected. The impedance of this external tuned grid driving circuit is very low at the doubling frequency, and thus there is no tendency for self-excited oscillation when ordinary triode tubes are used. At very high frequencies, however, this impedance may be great enough to cause regeneration, or even oscillation, at the tuned output frequency of the doubler.

A doubler can either be neutralized or made

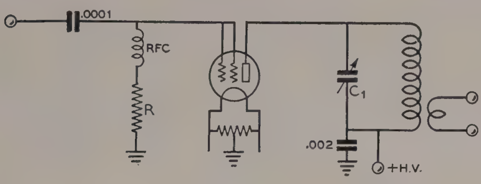


Figure 14.

CONVENTIONAL FREQUENCY DOUBLER CIRCUIT.

A high-μ, dual-grid triode, or a pentode or beam tetrode with the grid and screen paralleled, makes an excellent frequency doubler. In addition, all of these types of tubes have the advantage that when the excitation is removed their plate current will fall to a very low value. The plate circuit is tuned to twice the frequency of the exciting voltage on the grid.

more regenerative by adjusting C_2 in the circuit shown in Figure 15.

When condenser C_2 is of the proper value to neutralize the plate-to-grid capacity of the tube, the plate circuit can be tuned to twice the frequency (or to the same frequency) as that of the source of grid drive; the tube can be operated either as a neutralized amplifier or doubler. The capacity of C_2 can be increased so that the doubler will become *regenerative,* if the r.f. impedance of the external grid driving circuit is high enough at the output frequency of the stage.

Frequency doublers require bias of several times cutoff; high-μ tubes therefore are desirable for this type of service. Tubes which have amplification factors from 20 to 200 are suitable for doubler circuits. Tetrodes and pentodes usually have high amplification factors. Low-μ triodes, having amplification constants of from 3 to 10, are not applicable for doubler service because in some cases the grid voltage must be as high as the plate voltage for efficient doubling action. The necessary grid excitation voltage for high-μ tubes can be obtained more easily from average driver stages in conventional exciters.

Angle of Flow in Frequency Multipliers. The angle of plate current flow in a frequency multiplier is a very important factor in determining the efficiency. As the angle of flow is decreased for a given value of grid current, the efficiency increases. To reduce the angle of flow, higher grid bias is required so that the grid excitation voltage will exceed the cutoff value for a shorter portion of the exciting-voltage cycle. Frequency doublers of all types should have an angle of flow of 90 degrees or less, triplers 60 degrees or less, and quadruplers 45 degrees or less.

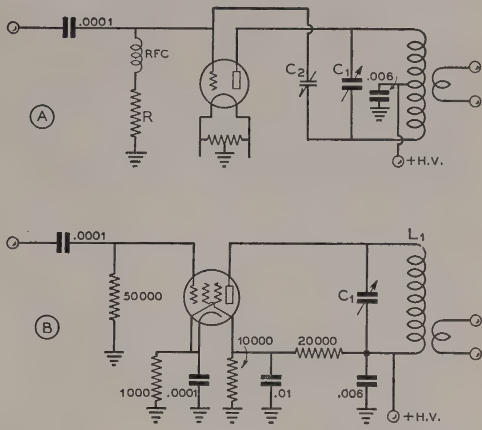


Figure 15.

REGENERATIVE DOUBLER CIRCUITS.

(A) shows a circuit which may be used either as a neutralized buffer stage or, when the capacity of C_2 is increased beyond the "neutralized" setting, as a regenerative doubler. (B) shows a frequency multiplier circuit with cathode regeneration which will give quite good results as a doubler, and very good results, compared to other multiplier circuits, as a frequency quadrupler.

Normally, a small angle of flow requires quite high bias and excitation. However, by altering the shape of the exciting voltage from its usual sine wave shape at the exciting frequency, it is possible to decrease the angle of flow and thus increase the efficiency without resorting to increases in the excitation voltage and bias.

The angle of flow may be decreased by adding some properly phased third harmonic voltage to the excitation. The result of adding the third harmonic voltage to the fundamental is shown graphically in Figure 16. As shown by the dotted curve, E_g, when the fundamental and third harmonic voltages are added in the proper phase, the result is a grid excitation voltage having a peaked wave form, exactly what is required for high-efficiency frequency multiplying. The method by which the third harmonic is added is shown in Figure 17. A small, center-tapped tank circuit tuned to three times the driver frequency is placed between the driver plate and the coupling condenser to the frequency-multiplier stage. The center tap of this coil is connected to the "hot" end of the driver plate tank, which remains tuned to the fundamental frequency. The third-harmonic tank circuit can be tuned

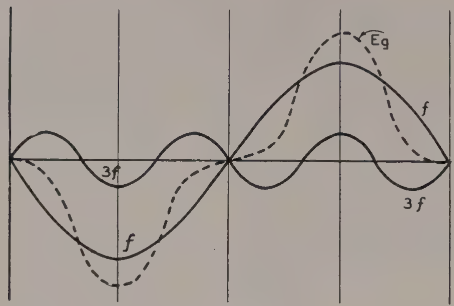


Figure 16.

PEAKED WAVEFORM OBTAINED BY ADDITION OF FUNDAMENTAL AND THIRD-HARMONIC ENERGY IN PROPER PHASE.

When fundamental frequency (f) energy and third-harmonic (3f) energy are added in the proper phase the result is a peaked waveform as shown by Eg. This peaked waveform, when used as excitation for a frequency doubler stage, gives considerably higher plate efficiency than when sine-wave excitation voltage is applied to the grid of the tube.

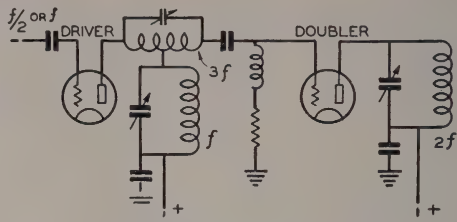


Figure 17.

CIRCUIT FOR COMBINING FUNDAMENTAL AND THIRD-HARMONIC ENERGY IN PROPER PHASE FOR PEAKED WAVEFORM.

The small third-harmonic tank circuit connected as shown adds the fundamental and third harmonic in the proper phase relation for producing a peaked excitation waveform on the grid of the doubler.

accurately to frequency by coupling to it a small, low-current dial lamp in a loop of wire and tuning for maximum brilliancy. An absorption wavemeter may be coupled to the third-harmonic tank after it has been tuned, to make sure that it is on the correct harmonic. The tuning of this circuit is not critical; one setting will serve to cover an amateur band.

Push-Push Doublers. Two tubes can be connected with the grids in push-pull, and the plates in parallel, for operation in a so-called *push-push doubler,* as shown in Figure 18.

This doubler circuit will deliver twice as much output as a single-tube circuit; it has proven popular in amateur transmitters because of its operating ease. In previous doubler circuits, capacitive coupling was shown. Link coupling to the tuned circuit in a preceding stage is shown in Figure 18. This coupling arrangement simplifies the push-pull connection of the two grid circuits.

The circuit C_2-L_2 is tuned to the same frequency as that of the preceding tuned circuit, and the doubler plate circuit C_1-L_1 is tuned to *twice* the frequency. The grid circuit should be tuned by means of a split-stator condenser, connected as shown in Figure 18, rather than by means of the single-section tuning condenser and by-passed center-tapped coil arrangement. The latter would provide a relatively high impedance at the doubling frequency. The push-push doubler then would be highly regenerative, and in most cases it would break into self-oscillation. The split-stator tuning circuit, because it has a capacitive reactance, provides a very low impedance at the doubling frequency, so that there is very little regenerative action; the circuit, therefore, is quite stable if the grid tank is not made too low C.

Tank Circuit Capacities

Tuning capacity values for class C amplifier tank circuits are an important consideration to anyone building a radio transmitter. The best value of capacity for a particular application can be determined closely by charts or formulas for any frequency of operation. The ratio of C to L, capacitance to inductance, depends upon the operating plate voltage and current, and upon the type of circuit. Proper choice of capacity-to-inductance ratio for resonance at any given frequency is important in obtaining low harmonic output, and also low distortion in the case of a modulated class C amplifier.

A class C amplifier produces a very distorted plate current wave form in the form of pulses as shown in Figure 19. The LC circuit is tuned to resonance, and its purpose is to smooth out these pulses, by its storage or "tank" action, into a sine wave of radio-frequency output. Any wave-form distortion of the carrier frequency is illegal, because it results in harmonic interference in higher-frequency channels. A class A radio-frequency amplifier would produce a sine wave output if the exciting voltage were a sine wave. However, the a.c. plate current would be flowing during the full 360° of each r.f. cycle, resulting

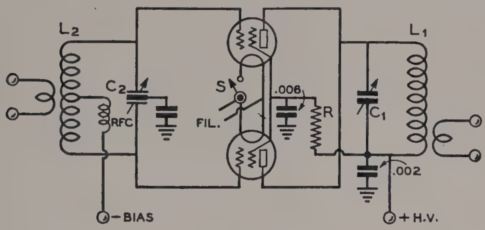
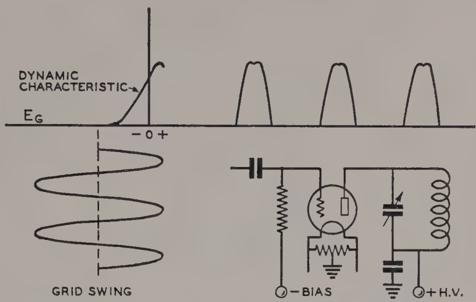


Figure 18.

PUSH-PUSH DOUBLER CIRCUIT.

In this type of doubler the grids are connected in push-pull and the plates are connected in parallel. A pair of triodes, a dual triode, or a pair of pentodes or tetrodes may be used. In the diagram shown, the heater of one of the tubes may be opened and the other tube operated as a neutralized amplifier, the other tube acting as the neutralizing condenser.



CLASS C AMPLIFIER PLATE CURRENT WAVEFORM

Figure 19.

in excessive plate loss in the tube for any reasonable value of output. The class C amplifier has a.c. plate current flowing during only a fraction of each cycle, allowing the plate to cool off during the remainder of the cycle. If the plate current is zero for 2/3 of each cycle, the *angle of plate current flow* is said to be 120°, since current is flowing during 1/3 of 360°. The tube in a class C amplifier could have several times as much power input for a given plate loss as when used in a class A amplifier.

The tuned circuit must have a good flywheel effect in order to furnish a sine-wave output to the antenna when it is receiving energy in the form of very distorted pulses such as shown in Figure 19. The LC circuit fills in power over the complete r.f. cycle, providing the LC ratio is correct. The flywheel effect is generally defined as the ratio of radio-frequency volt-amperes to actual power output, or VA/W. This is equivalent to Q and should not be much less than 4π, or 12.5, for a class-C amplifier. At this value of VA/W or Q, one-half of the stored energy in the LC circuit is absorbed by the antenna. If a lower value of Q is used, the storage power is insufficient to produce a sine (undistorted) wave output to the antenna and power will be wasted in radiation of harmonics.

Too high a value of VA/W or Q will result in excessive circulating r.f. current loss in the LC circuit and lowered output to the antenna. In high-fidelity radiophone transmitters, too high a Q will cause attenuation of the higher sideband frequencies and consequent loss of the higher audio frequencies.

Harmonic Radiation vs. Q. Opinions vary as to the correct value of Q, but a careful analysis of the whole problem seems to indicate that a value of 12 is suitable for most amateur phone or c.w. transmitters. A value of 15 to 20 will result in less harmonic radiation at the expense of a little additional heat power loss in the tank or LC circuit. The charts shown have been calculated for an operating value of Q = 12.

The curves shown in Figure 20 indicate the sharp increase in harmonic output into the antenna circuit for low values of Q. The curve for the second harmonic rises nearly vertically for Q values of less than 10. The third harmonic does not become seriously large for values of Q less than 4 or 5. These curves show that push-pull amplifiers may be operated at lower values of Q if necessary, since the second harmonic is cancelled to a large extent *if there is no capacitive or unbalanced coupling* between the tank circuit and the antenna feeder system.

Effect of Loading on Q. The Q of a circuit depends upon the resistance in series with the capacitance and inductance. This

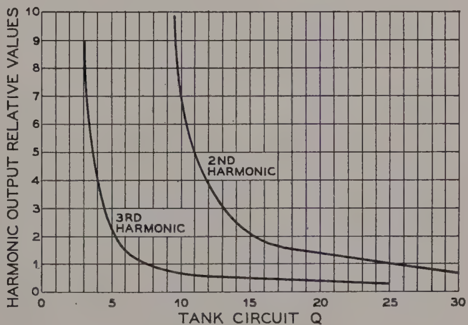
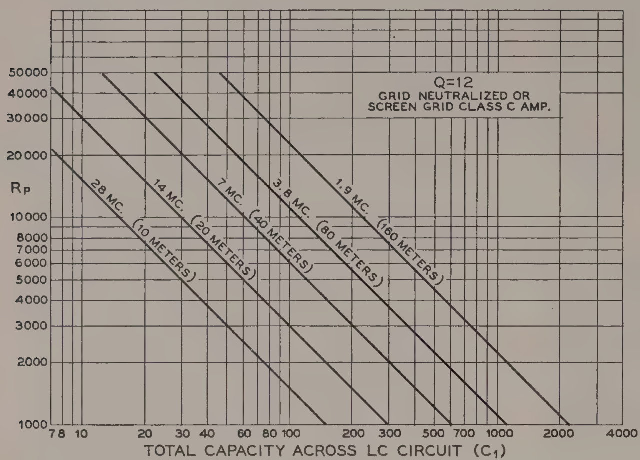
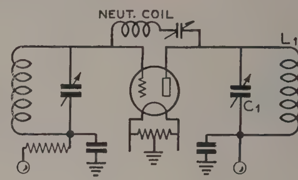


Figure 20.

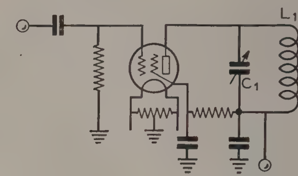
SECOND AND THIRD HARMONIC OUTPUT PLOTTED AGAINST TANK CIRCUIT Q.



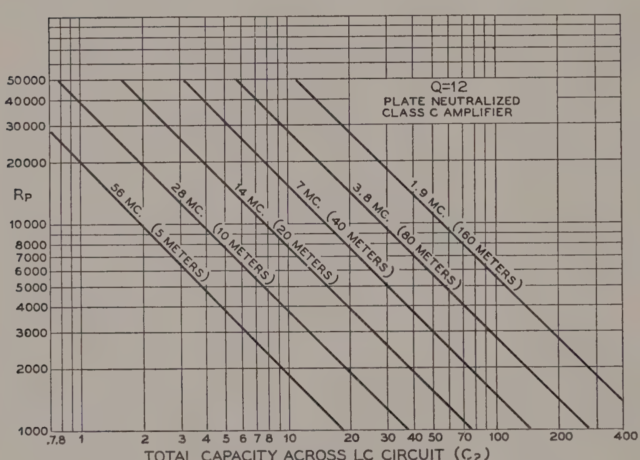


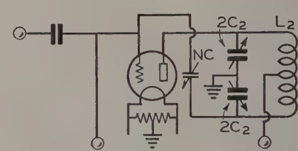
(A) COIL NEUTRALIZED AMPLIFIER

Figure 21.



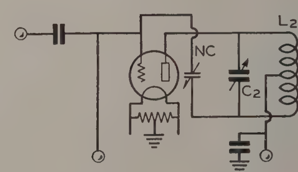
(B) SCREEN GRID AMPLIFIER





(A)

Figure 22.



(B) PLATE NEUTRALIZED AMPLIFIERS

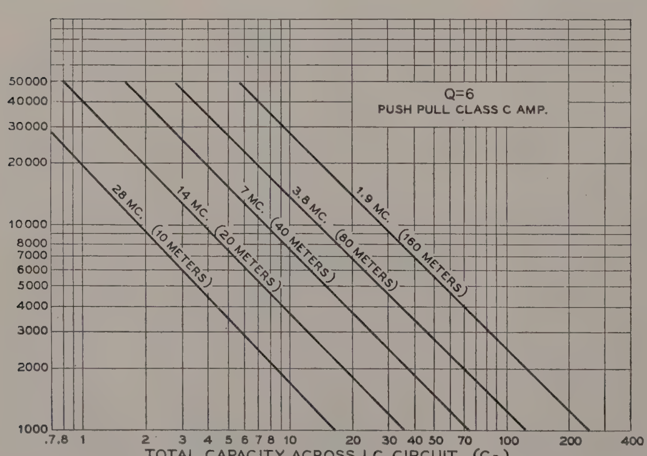
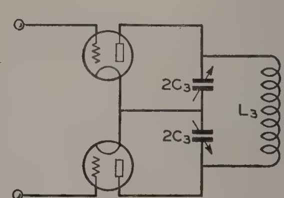


Figure 23.



PUSH PULL AMPLIFIER

series resistance is very low for a low-loss coil not loaded by an antenna circuit. The value of Q may be from 100 to 300 under these conditions. Coupling an antenna circuit has the effect of increasing the series resistance, though in this case the power is consumed as useful radiation by the antenna. Mathematically, the antenna increases the value of R in the expression $Q = \omega L/R$ where L is the coil inductance and ω is the term $2\pi f$, f being in cycles per second.

The antenna coupling can be varied to obtain any value of Q from 3 to values as high as 100 or 200. However, the value of Q = 12 (or Q = 20 if desired) will not be obtained at normal values of d.c. plate current in the class C amplifier tube unless the C-to-L ratio in the tank circuit is correct for that frequency of operation.

The values of C_1, C_2 and C_3 shown in Figures 21, 22 and 23 are for the total capacity across the inductance. This includes the tube interelectrode capacities, distributed coil capacity, wiring capacities and tuning condenser capacity. If a split-stator condenser is used, the effective capacity is equal to half of the value of each section, since the two sections are in series across the tuned circuit. The total stray capacities range from approximately 2 up to 30 $\mu\mu$fd. and largely depend upon the type of tube or tubes used in the class C amplifier.

In the push-pull circuit of Figure 23, each tube works on a portion of each half cycle, so less storage or flywheel effect is needed and a value of Q = 6 may be used instead of Q = 12.

The values of R_p are easily calculated by dividing the d.c. plate supply voltage by the total d.c. plate current (expressed in amperes). Correct values of total tuning capacity are shown in the charts for the different amateur bands. The shunt stray capacity can be estimated closely enough for all practical purposes. The coil inductance should then be chosen which will produce resonance at the desired frequency with the total calculated tuning capacity.

The capacities shown are the minimum recommended values and they should be increased 50 per cent to 100 per cent for modulated class C amplifiers where economically feasible. The values shown in the charts are sufficient for c.w. operation of class C amplifiers. It is again emphasized that these values are *total capacities* across the tank circuit, and should not be considered as the capacity *per section* for a *split-stator* condenser. If a split-stator condenser is to be used, the *per section* capacity should be *twice* that indicated by the charts shown on the opposite page.

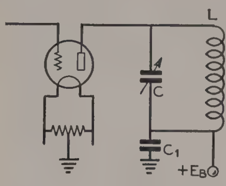


Figure 24.

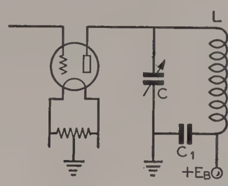


Figure 25.

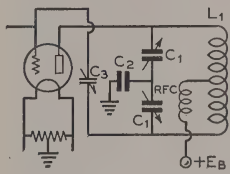


Figure 26.

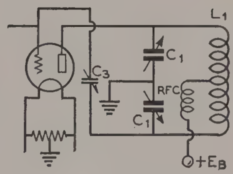


Figure 27.

Tuning Condenser Air Gap

Plate-Spacing Requirements for Various Circuits and Plate Voltages. In determining condenser air gaps, the peak r.f. voltage impressed across the condenser is the important item, since the experimental and practical curves of air gap versus peak volts as published by the Allen D. Cardwell Mfg. Corp. may be applied to any condenser with polished plates having rounded edges. Typical peak breakdown voltages for corresponding air gaps are listed in the table. These values can be used in any circuit. The problem is to find the peak r.f. voltage in each case; this can be done quite easily.

The r.f. voltage in the plate circuit of a class C amplifier tube varies from nearly zero to twice the d.c. plate voltage. If the d.c. voltage is being 100 per cent modulated by an audio voltage, the r.f. peaks will reach four times the d.c. voltage. These are the highest values reached in any type of loaded amplifier: a class B linear, class C grid- or plate-modulated, or class C c.w. amplifier. The circuits shown in Figures 25 and 27 require a tuning condenser with plate spacing which will have an r.f. peak breakdown rating at least equal to 2 times or 4 times the d.c. plate voltage for c.w. and plate-modulated amplifiers respectively.

It is possible to reduce the air gap to one-half by connecting the amplifier so that the d.c. plate voltage does not appear across the tuning condenser. This is done in Figures 24 and 26. These circuits should always be used in preference to those of Figures 25 and

27, since the tuning condenser is only about one-fourth as large physically for the same capacity. Consequently, it is proportionately less expensive.

The peak r.f. voltage of a plate-modulated class C amplifier varies at 100 per cent modulation from nearly zero to four times E_b, the d.c. plate voltage, but only one-half of this voltage is applied across the tuning condensers of Figures 24 and 26. For a class B linear, class C grid-modulated, or c.w. amplifier, the r.f. voltage across the tube varies from nearly zero up to twice E_b. The r.f. voltage is an a.c. voltage varying from zero to a positive and then to a negative maximum over each cycle. The fixed (mica) condenser C_1 in Figure 24, and C_2 in Figure 26 insulates the rotor from d.c. and allows us to subtract the d.c. voltage value from the tube peak r.f. voltage value in calculating the breakdown voltage to be expected.

This gives us a simple rule to follow for a normally-loaded plate-modulated r.f. amplifier. The peak voltage across the tuning condenser C or C_1 of Figures 24 and 26, respectively, will be *twice the d.c. plate voltage.* If a single-section condenser is used in Figure 26, with the by-pass condenser C_2 connected to the coil center tap, the plate spacing or air gap must be twice as great as that of a split-stator condenser; so there is no appreciable saving in costs for a given capacity.

In c.w. amplifiers the air gap must be great enough to withstand a peak r.f. voltage *equal to the d.c. plate voltage,* for each section C_1 of Figure 26, or, C of Figure 26.

These rules apply to a loaded amplifier or buffer stage. If the latter is ever operated without an r.f. load, the peak voltages may be very much greater—by as much as two or three times in ordinary LC circuits. For this reason no amplifier should be operated without load when anywhere near normal d.c. plate voltage is applied.

A factor of safety in the air-gap rating should be applied to insure freedom from r.f. flashover. This is especially true when using the circuits of Figures 25 and 27; in these circuits the plate supply is shorted when a flashover occurs. Knowing the peak r.f. voltage, an air gap should be chosen which will be about 100 per cent greater than the breakdown rating. The air gaps listed will break down at the approximate peak voltages in the table. If the circuits are of the form shown in Figures 25 and 27, the peak voltages across the condensers will be nearly twice as high, and twice as large an air gap is needed. The fixed condensers, usually of the mica type, shown in Figures 24 and 26, must be rated to withstand the d.c. plate voltage plus any audio voltage. This condenser should be rated at a d.c. working voltage of at least *twice the d.c. plate supply in a plate modulated amplifier,* and at least *equal to the d.c. supply* in any other type of r.f. amplifier.

Push-Pull Stages. The circuits of Figures 26 and 27 apply without any change in calculations to push-pull amplifiers. Only one tube is supplying power to the tuned circuit at any given instant, each one driving a part of each half cycle. The different value of Q and increased power output increase the peak voltages slightly, but, for all practical purposes, the same calculation rules may be employed.

These rules are based on average amateur design for any form of r.f. amplifier, with a recommended factor of safety of 100 per cent to prevent flashover in the condenser.

BREAKDOWN RATINGS OF COMMON PLATE SPACINGS

AIR-GAP IN INCHES	PEAK VOLTAGE BREAKDOWN
.030	1000
.050	1500
.070	3000
.078	3500
.084	3800
.100	4150
.144	5000
.175	5700
.200	6200
.250	7200
.300	8200
.350	9250
.375	10,000
.500	12,000

Recommended Air gap (approx. 100% factor of safety) for the circuits of figures 24 and 26. Spacings should be multiplied by 1.5 for same factor of safety with circuits of Figures 25 and 27.

D.C. PLATE VOLTAGE	C. W.	PLATE MOD.
400	.030	.050
600	.050	.070
750	.050	.100
1000	.070	.084
1250	.070	.144
1500	.078	.200
2000	.100	.250
2500	.175	.375
3000	.200	.500
3500	.250	.600

This is sufficient for operation into normal loads at all times, providing there are no freak parasitic oscillations present. The latter sometimes cause flashover across air gaps which should ordinarily stand several times the normal peak r.f. voltages. This is especially true of low-frequency parasitics.

The actual peak voltage values of a stable, loaded r.f. amplifier are somewhat less than the calculations indicate, which gives an additional factor of safety in the design.

Parasitic Oscillation in R.F. Amplifiers

Parasitics are undesirable oscillations either of very high or very low frequencies which occur in radio-frequency amplifiers.

They may cause additional signals (which are often rough in tone), other than normal harmonics, hash on each side of a modulated carrier, key clicks, voltage breakdown or flashover, instability or inefficiency, and shortened life or failure of the tubes. They may be damped and stop by themselves after keying or on modulation cycles, or they may be undamped and built up during ordinary unmodulated transmission, continuing if the excitation is removed. They may be at audio or radio frequency, in either type of amplifier (though only the r.f. amplifier is treated in this discussion). They may result from series or parallel resonant circuits of all types. Due to the neutralizing lead length or the nature of most parasitic circuits, the amplifier usually is not neutralized for the parasitic frequency.

Sometimes the fact that the plate supply is keyed obscures parasitic oscillations that might be very severe if the plate voltage were left on and only the excitation removed.

In some cases, an all-wave receiver will prove helpful in finding out if the amplifier is without spurious oscillations, but it may be necessary to check from one meter on up, to be perfectly sure. A normal harmonic is weaker than the fundamental but of good tone; a strong harmonic or a rough note at any frequency generally indicates trouble.

Low-Frequency Parasitics. One type of unwanted oscillation often occurs in shunt-fed circuits in which the grid and plate chokes resonate, coupled through the tube's inter-electrode capacity. It can also happen with series feed. This oscillation is generally at a much lower frequency than the desired one and causes additional carriers to appear, spaced from twenty to a few hundred kilocycles on either side of the main wave. One cure is to change the type of feed in either the grid or plate circuit or to eliminate one choke. Another is to use much less inductance in the grid choke than in the plate choke, or to replace the grid choke by a wire-wound resistor if the grid is series fed. In a class C stage with grid-leak bias, no r.f. choke is required if the bias is series fed.

This type of parasitic may take place in push-pull circuits, in which case the tubes are effectively in parallel for the parasitic and hence, the neutralization is not effective. The grids or plates can be connected together without affecting the undesired oscillation; this is a simple test for this type of parasitic.

Parallel Tubes. A very high frequency inter-tube oscillation often occurs when tubes are operated in parallel. Noninductive damping resistors or manufactured parasitic suppressors in the grid circuit, or short interconnecting grid leads, together with small plate choke coils, very likely will prove helpful.

Tapped Inductances. When capacity coupling is used between stages, particularly when one of the stages is tapped down from the end of the coil, additional parasitic circuits are formed because of the multiple resonant effects of this complex circuit. Inductive or link coupling permits making adjustments without forming these undesired circuits. Likewise, a condenser tapped across only part of an inductance, for bandspread tuning or capacity loading, can give rise to parasitics.

Multi-Element Tubes. Screen-grid, pentode, and beam tetrode tubes may help to eliminate parasitic circuits by using no neutralization, but their high gain occasionally causes parasitic oscillations. Furthermore, the by-pass circuit from the additional elements to the filament must be short and effective, particularly at the higher frequencies, to prevent undesired internal coupling. At the high frequencies, a variable screen by-pass condenser at some settings may improve the internal shielding without causing a new parasitic oscillation. A blocking (relaxation) effect may occur if the screen is fed through a series resistor. The screen circuit can, of course, act as the plate in a tuned-grid tuned-plate oscillation that can be detuned or damped at the control grid terminal.

Crystal Stages. Crystal oscillators are seldom suspected of parasitic oscillation troubles, but are often guilty. Ordinary as well as parasitic circuit coupling between the grid and plate circuits should be held to a minimum by separating the grid and plate leads, and by reducing the area of the loop from the grid through the crystal holder to the filament. Keeping the grid circuit short, even adding a small choke coil of a few turns in the plate lead next to the tube, will probably eliminate the possibility of high-voltage series-tuned parasitic oscillations.

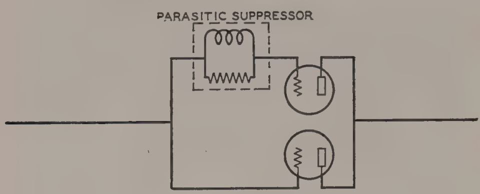


Figure 28.

Showing the use of a parasitic suppressor in series with one grid of a pair of paralleled tubes. In a push-pull amplifier which develops parasitics, the parasitic suppressor can be connected in series with the lead from the grid tank circuit to the grid of one of the tubes.

Parasitic Suppressors. The most common type of parasitic is of the u.h.f. type, which fortunately can usually be dampened by inserting a parasitic suppressor of the type illustrated in Figure 28 in the grid lead, or in *one* grid lead of either a push-pull or parallel tube amplifier.

Grid Bias

Radio-frequency amplifiers require some form of *grid bias* for proper operation. Practically all r.f. amplifiers operate in such a manner that plate current flows in the form of short peaked impulses which have a duration of only a fraction of an r.f. cycle. The plate current is cut off during the greater part of the r.f. cycle, which makes for high efficiency and high power output from the tubes, since there is no power being dissipated by the plates during a major portion of each r.f. cycle. The grid bias must be sufficient to cut off the plate current, and in very high efficiency class C amplifiers this bias may be several times the cutoff value. Cutoff bias, it will be recalled, is that value of grid voltage which will reduce the plate current to zero. The method for calculating it has been indicated previously. This theoretical value of cutoff will not reduce the plate current completely to zero, due to the variable-μ tendency which is characteristic of all tubes as the cutoff point is approached. This factor, however, is of no importance in practical applications.

Class C Bias. Radiophone class C amplifiers should be operated with the grid bias adjusted to values between two and three times cutoff at normal values of d.c. grid current to permit linear operation (necessary when the stage is plate-modulated). C.w. telegraph transmitters can be operated with bias as low as cutoff, if limited excitation is available and high plate efficiency is not a factor. In a c.w. transmitter, the bias supply or resistor should be adjusted to the point which will allow normal grid current to flow for the particular amount of grid driving r.f. power available. This form of adjustment will allow more output from the under-excited r.f. amplifier than when twice cutoff, or higher bias is used with low values of grid current.

Grid-Leak Bias. A resistor can be connected in the grid circuit of an r.f. amplifier to provide grid-leak bias. This resistor R_1 in Figure 29 is part of the d.c. path in the grid circuit.

The r.f. excitation is applied to the grid circuit of the tube. This causes a pulsating d.c. current to flow through the bias supply lead and any current flowing through R_1 produces a voltage drop across that resistance. The grid of the tube is positive for a short duration of each r.f. cycle, and draws electrons from the filament or cathode of the tube during that time. These electrons complete the circuit through the d.c. *grid return.* The voltage drop across the resistance in the grid return provides a *negative bias* for the grid. The r.f. chokes in Figures 29, 30, 31, and 32 prevent the r.f. excitation from flowing through the bias supply, or from being short-circuited to ground. The by-pass condenser across the bias source proper is for the purpose of providing a low impedance path for the small amount of stray r.f. energy which passes through the r.f. choke.

Grid-leak bias automatically adjusts itself even with fairly wide variations of r.f. excitation. The value of grid-leak resistance should be such that normal values of grid current will flow at the maximum available amount of r.f. excitation. Grid-leak bias cannot be used for grid-modulated or linear amplifiers in which the average d.c. grid current is constantly varying with modulation.

Grid-leak bias alone provides no protection against excessive plate current in case of failure of the crystal oscillator, or failure of any other source of r.f. grid excitation. A *C-battery* or *C-bias* supply can be connected in series with the grid leak, as shown in Figure 30. This additional C-bias should be made at least equal to cutoff bias. This will protect the tube in the event of failure of grid excitation. Zero-bias tubes do not require this bias source in addition to the grid leak, since their plate current will drop to a safe value when the excitation is removed.

Cathode Bias. A resistor can be connected in series with the cathode or center-tapped filament lead of an amplifier to secure *automatic bias*. The plate current flows through this resistor, then back to the cathode or fila-

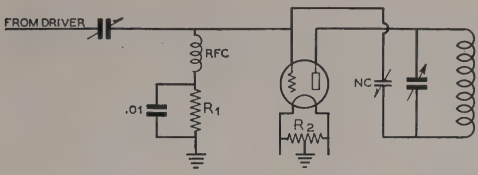


Figure 29.
GRID LEAK BIASED STAGE.
Showing how a resistor may be connected in series with the grid return lead to obtain bias due to the flow of rectified grid current through the resistor.

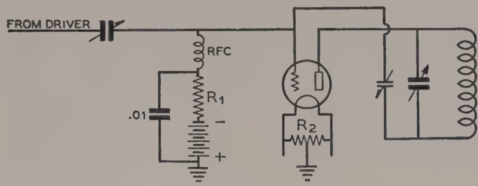


Figure 30.
GRID LEAK AND BATTERY BIAS.
A battery may be added to the grid-leak bias system of Figure 30 to provide protection in case of excitation failure.

ment, and the voltage drop across the resistor can be applied to the grid circuit by connecting the grid bias lead to the grounded, or power supply end of the resistance R, as shown in Figure 31.

The grounded (B-minus) end of the cathode resistor is negative relative to the filament by an amount equal to the voltage drop across the resistor. The value of resistance must be so chosen that the sum of the desired grid and plate current flowing through the resistor will bias the tube for proper operation.

This type of bias is used more extensively in audio-frequency than in radio-frequency amplifiers. The voltage drop across the resistor must be subtracted from the total plate supply voltage when calculating the power input to the amplifier, and this loss of plate voltage in an r.f. amplifier may be excessive. A class A audio amplifier is biased only to approximately one-half cutoff, whereas an r.f. amplifier may be biased to twice cutoff, or more, and thus the plate supply voltage loss may be a large percentage of the total available voltage when using low- or medium-μ tubes.

Separate Bias Supply. C-batteries, or an external C-bias supply, sometimes are used for grid bias to an amplifier, as shown in Figure 32.

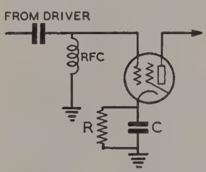
Battery bias gives very good voltage regulation and is satisfactory for grid-modulated or linear amplifiers, which operate nearly at zero grid current. In the case of class C amplifiers which operate with high grid current, battery bias is not very satisfactory. This direct current has a charging effect on the dry batteries; after a few months of service the cells will become unstable, bloated, and noisy.

A separate a.c. operated power supply can be used as a substitute for dry batteries. The bleeder resistance across the output of the filter can be made sufficiently low in value that the grid current of the amplifier will not appreciably change the amount of negative grid-bias voltage. This type of bias supply is used in class B audio and class B r.f. linear amplifier service where the voltage regulation in the C-bias supply is important. For a class C amplifier it is not so important, and an economical design of components in the power supply, therefore, can be utilized. However, in a class C application the bias voltage must be adjusted with normal grid current flowing as the grid current will raise the bias when it is flowing through the bias-supply bleeder resistance.

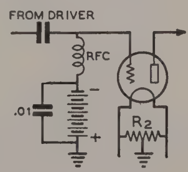
Interstage Coupling

Energy is usually coupled from one circuit in a transmitter into another in the following ways: *capacitive coupling, inductive coupling,* or *link coupling.* The latter is a special form of inductive coupling. The choice of a coupling method depends upon the purpose for which it is to be used.

Capacitive Coupling. Capacitive coupling between an amplifier or doubler circuit and a



CATHODE BIAS
FIGURE 31



BATTERY BIAS
FIGURE 32

A resistor in the cathode lead gives cathode, or "automatic" bias as shown in Figure 31. The voltage drop across the cathode resistor due to the flow of plate and grid current is applied to the grid in the form of negative bias. Figure 32 shows the use of a battery only as bias—this arrangement is suitable only for stages which do not draw over about 15 ma. of grid current.

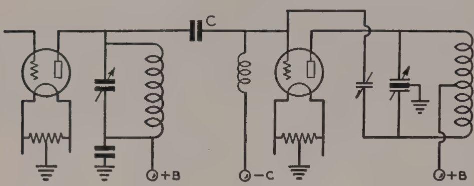


Figure 33.
CAPACITIVE INTERSTAGE COUPLING.

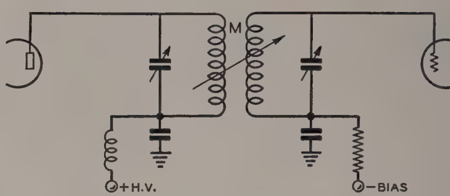


Figure 35.
INDUCTIVE INTERSTAGE COUPLING.

preceding driver stage is shown in Figure 33.

The coupling condenser, C, isolates the d.c. plate supply from the next grid and provides a low impedance path from the r.f. energy between the tube being driven and the driver tube. This method of coupling is simple and economical for low-power amplifier or exciter stages, but has certain disadvantages. The grid leads in an amplifier should be as short as possible, but this is difficult to attain in the physical arrangement of a high-power amplifier with respect to a capacitively-coupled driver stage.

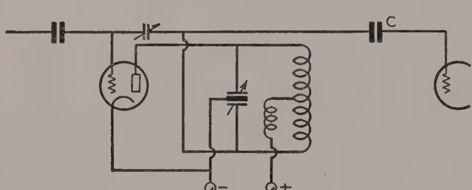


Figure 34.

BALANCED CAPACITIVE COUPLING.

This type of capacitive interstage coupling helps to equalize the capacities across the two sides of the driver tank circuit.

Disadvantages of Capacity Coupling. The r.f. choke in series with the C-bias supply lead must offer an extremely high impedance to the r.f. circuit, and this is difficult to obtain when the transmitter is operated on several harmonically related bands. Another disadvantage of capacitive coupling is the difficulty of adjusting the load on the driver stage. Impedance adjustment can be accomplished by tapping the coupling lead a part of the way down on the plate coil of the tuned stage of the driver circuit. However, when this lead is tapped part way down on the coil, a parasitic oscillation tendency becomes very troublesome and is difficult to eliminate. If the driver stage has sufficient power output so that an impedance mismatch can be tolerated, the condenser C in Figure 33 can be connected directly to the top of the coil, and made small enough in capacity for the particular frequency of operation that not more than normal plate current is drawn by the driver stage.

The grid circuit impedance of a class C amplifier may be as low as a few hundred ohms in the case of a high-μ tube, and may range from that value up to a few thousand ohms for low-μ tubes.

Capacitive coupling places the grid-to-filament capacity of the driven tube directly across the driver tuned circuit, which reduces the LC ratio and sometimes makes the r.f. amplifier difficult to neutralize because the additional driver stage circuit capacities are connected into the grid circuit. Difficulties from this source can be partially eliminated by using a center-tapped or split-stator tank circuit in the plate of the driver stage, and capacity coupling to the opposite end from the plate. This method places the plate-to-filament capacity of the driver across one-half of the tank and the grid-to-filament capacity of the following stage across the other half. This type of coupling is shown in Figure 34.

Capacitive coupling can be used to advantage in reducing the total number of tuned circuits in a transmitter so as to conserve space and cost. It also can be used to advantage between stages for driving tetrode or pentode amplifier or doubler stages. These tubes require relatively small amounts of grid excitation.

Inductive Coupling. The r.f. amplifier often is coupled to the antenna circuit by means of *inductive coupling,* which consists of two coils electromagnetically coupled to each other. The antenna tuned circuit can be of the series-tuned type, such as is illustrated for *Marconi*-type 160-meter antennas in the chapter on *Antennas.* Parallel resonant circuits sometimes are used, as shown in Figure 35, in which the antenna feeders are connected across

the whole or part of the secondary circuit.

The degree of coupling is controlled by varying the mutual inductance of the two coils, which is accomplished by changing the spacing between the coils.

Inductive coupling also is used extensively for coupling r.f. amplifiers in radio receivers, and occasionally in transmitting r.f. amplifier circuits. The mechanical problems involved in adjusting the degree of coupling in a transmitter make this system of limited practical value.

Link Coupling. A special form of inductive coupling which is applied to radio transmitter circuits is known as *link coupling.* A low impedance r.f. transmission line, commonly known as a *link,* couples the two tuned circuits together. Each end of the line is terminated in one or more turns of wire, or *loops,* wound around the coils which are being coupled together. These loops should be coupled to each tuned circuit at the point of zero r.f. potential. This *nodal* point is the center of the tuned circuit in the case of plate-neutralized or push-pull amplifiers, and at the positive-B end of the tuned circuit in the case of screen grid and Hazeltine-neutralized amplifiers.

The nodal point in an antenna tuned circuit depends upon the type of feeders, and the node may be either at the center or at one end of the tuned circuit.

The nodal point in tuned grid circuits is at the C-bias or grounded end of plate-neutralized or screen-grid r.f. amplifiers, and at the center of the tuned grid coil in the case of push-pull amplifiers. The link coupling turns should be as close to the nodal point as possible. A ground connection to one side of the link is used in special cases where harmonic elimination is important, or where capacitive coupling between two circuits must be minimized.

Typical link coupled circuits are shown in Figures 36 and 37.

Some of the advantages of link coupling are listed here:

(1) It eliminates coupling taps on tuned circuits.

(2) It permits the use of series power supply connections in both tuned grid and tuned plate circuits, and thereby eliminates the need of r.f. chokes.

(3) It allows separation between transmitter stages of distances up to several feet without appreciable r.f. losses.

(4) It reduces capacitive coupling and thereby makes neutralization more easily attainable in r.f. amplifiers.

(5) It provides semi-automatic impedance matching between plate and grid tuned circuits, with the result that greater grid swing can be obtained in comparison to capacitive coupling.

(6) It effectively reduces harmonic radiation when a final amplifier is coupled to a tuned antenna circuit, due to the additional tuned circuit and, particularly, due to the reduction of capacitive coupling to the antenna.

The link coupling line and loops can be made of no. 18 or 20 gauge push-back wire for coupling low-power stages. High-power circuits can be link-coupled by means of no. 8 to no. 12 rubber-covered wire, twisted low-impedance antenna-feeder wire, concentric lines or open-wire lines of no. 12 or no. 14 wire spaced ¼ to ½ inch.

The impedance of a link coupling line varies from 75 to 200 ohms, depending upon the diameter of the conductors and the spacing between them.

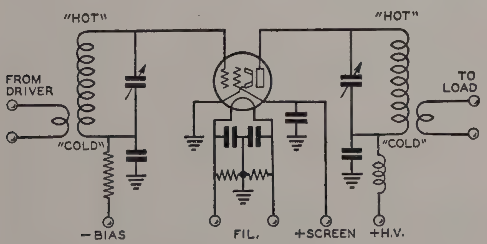


Figure 36.
LINK COUPLED CIRCUIT.
Showing link coupling into and out of a single-ended beam-tetrode amplifier stage. The coupling links should be placed at the "cold" or low-potential ends of the grid and plate coils.

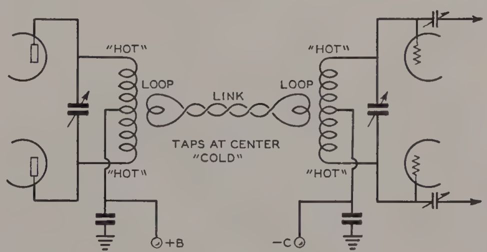
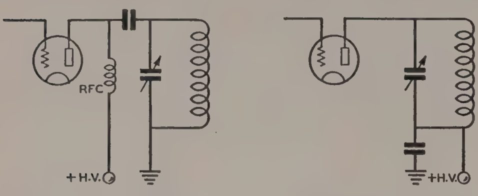
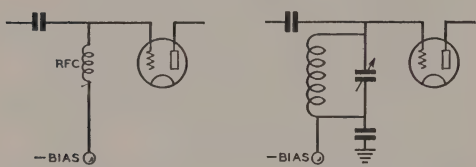


Figure 37.
PUSH-PULL LINK COUPLING.
When link coupling is used between push-pull stages or between "split" tank circuits, the coupling loops are placed at the center of the coils.



PARALLEL PLATE FEED SERIES PLATE FEED

Figure 38.



SHUNT BIAS FEED SERIES BIAS FEED

Figure 39.

Radio-Frequency Chokes

Radio-frequency chokes are connected in circuits for the purpose of preventing r.f. energy from being short-circuited, or escaping into power supply circuits. They consist of inductances wound with a large number of turns, either in the form of a solenoid or universal pie-winding. These inductances are designed to have as much inductance and as little distributed or shunt capacity as possible, since the capacity by-passes r.f. energy. The unavoidable small amount of distributed capacity resonates the inductance, and this frequency normally should be lower than the frequency at which the transmitter or receiver circuit is operating. R.f. chokes for operation on several harmonically related bands must be designed carefully so that the impedance of the choke will be extremely high (several hundred thousand ohms) in each of the bands.

The direct current which flows through the r.f. choke largely determines the size of wire to be used in the windings. The inductance of r.f. chokes for very short wave-lengths is much less than for chokes designed for broadcast and ordinary short-wave operation, so that the impedance will be as high as possible in the desired range of operation. A very high inductance r.f. choke has more distributed capacity than a smaller one, with the result that it will actually offer *less* impedance at very high frequencies.

Shunt and Series Feed. Direct-current grid and plate connections are made either by *series* or *parallel feed* systems. Simplified forms of each are shown in Figures 38 and 39.

Series feed can be defined as that in which the d.c. connection is made to the grid or plate circuit at a point of very low r.f. potential. Shunt feed always is made to a point of high r.f. voltage and always requires a high impedance r.f. choke or resistance in the connection to the high r.f. point to prevent loss of r.f. power.

Parallel and Push-Pull Tube Circuits

The comparative r.f. power output from parallel or push-pull operated amplifiers is the same if proper impedance matching is accomplished, if sufficient grid excitation is available in both cases, and if the frequency of measurement is considerably lower than the frequency limit of the tubes.

Parallel Operation. Operating tubes in parallel has some advantages in transmitters designed for operation on 40, 80, and 160 meters, or for broadcast band operation. Only one neutralizing condenser is required for parallel operation, as against two for push-pull. However, on wavelengths below 40 meters, parallel tube operation is not advisable because of the unbalance in capacity across the tank circuits. Low-C types of vacuum tubes can be connected in parallel with less difficulty than the high-C types, in which the combined interelectrode capacities might be quite high in the parallel connection.

Push-Pull Operation. The push-pull connection provides a well-balanced circuit insofar as miscellaneous capacities are concerned; in addition, the circuit can be neutralized more easily, especially in high-frequency amplifiers. The L/C ratio in a push-pull amplifier can be made higher than in a plate-neutralized parallel-tube operated amplifier. Push-pull amplifiers, when perfectly balanced, have less second-harmonic output than parallel or single-tube amplifiers. In actual practice, however, undesired capacitive coupling and circuit unbalance tend to offset part of the theoretical harmonic-reducing advantage of push-pull r.f. circuits.

Transmitter Keying

The carrier frequency signal from a c.w. transmitter must be broken into dots and dashes in the form of *keying* for the transmission of code characters. The carrier signal is of a constant amplitude while the key is closed, and is entirely removed when the key is open. If the change from the no-output condition to *full-output* occurs too rapidly, an undesired *key-click* effect takes place which causes interference in other signal channels. If the opposite condition of full output to no

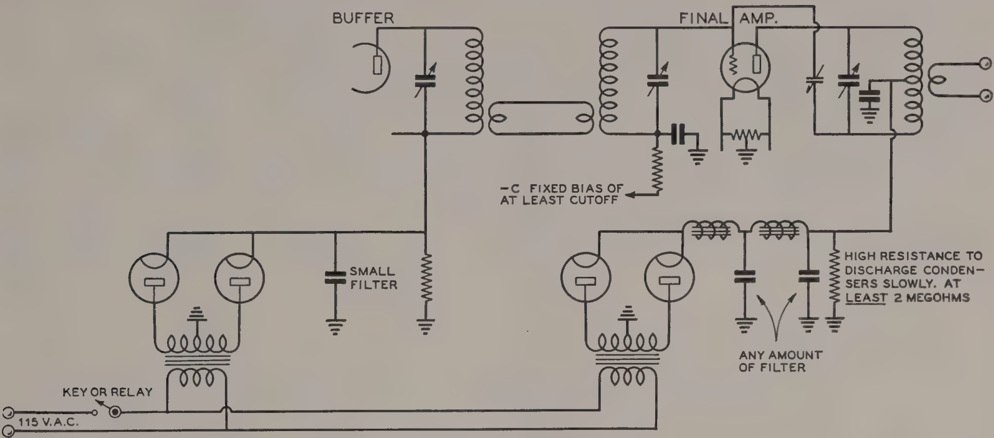


Figure 40.
IMPROVED PRIMARY KEYING WITHOUT CLICKS OR "TAILS."

output condition occurs too rapidly, a similar effect takes place.

Excitation or Plate Voltage Keying. The two general methods of keying a c.w. transmitter are those which control either the excitation, or the plate voltage which is applied to the final amplifier. Plate voltage control can be obtained by connecting the key in the primary line circuit of the high voltage plate power supply. A slight modification of direct plate voltage control is the connection of the c.w. key or relay in the filament center-tap lead of the final amplifier. *Excitation keying* can be of several forms, such as crystal oscillator keying, buffer stage keying, or blocked-grid keying.

Key Clicks. Key clicks should be eliminated in all c.w. telegraph transmitters. Their elimination is accomplished by preventing a too-rapid make-and-break of power to the antenna circuit. A gradual application of power to the antenna, and a similarly slow cessation, will eliminate key clicks. Too much lag will prevent fast keying, but fortunately key clicks can be practically eliminated without limiting the speed of manual (hand) keying. Some circuits which eliminate key clicks introduce too much time-lag and thereby add *tails* to the dots. These tails may cause the signals to be difficult to copy at high speeds.

Click Filters. Eliminating key clicks by some of the key-click filter circuits illustrated in the following text is not certain with every individual transmitter. The constants in the time-lag and spark-producing circuits depend upon the individual characteristics of the transmitter, such as the type of filter, power input, and various circuit impedances. All keying systems have one or more disadvantages, so that no particular method can be recommended as an ideal one. An intelligent choice can be made by the reader for his particular transmitter requirements by carefully analyzing the various keying circuits.

Primary Keying. Key clicks (except those arising from arcing at the key, which usually do not carry beyond a few hundred feet) can be eliminated entirely by means of primary keying, in which the key is placed in the a.c. line supply to the primary of the high voltage plate supply transformer. This method of keying also has the advantage that grid leak bias can be used in the keyed stages of the transmitter. As ordinarily applied, the plate voltage to the final amplifier is controlled by the action of the key. The filter in the high voltage rectifier circuit creates a time lag in the application and removal of the d.c. power input to the r.f. amplifier. Too much filter will introduce too great a time lag, and add tails to the dots.

A heavy-duty key or keying relay is necessary for moderate or high-power transmitters to break the inductive a.c. load of the power supply. The exciting current or surge current may be several times as high as the average current drawn by the transformer which is being keyed. This will cause difficulty from sticking key contacts or burnt points on the keying relay. This effect can be minimized by proper design of the power transformer, which should have a high primary inductance and an iron core of generous size.

Lag-Less Primary Keying Circuit. An improved primary keying circuit is shown in Figure 40. This circuit makes high speed key-

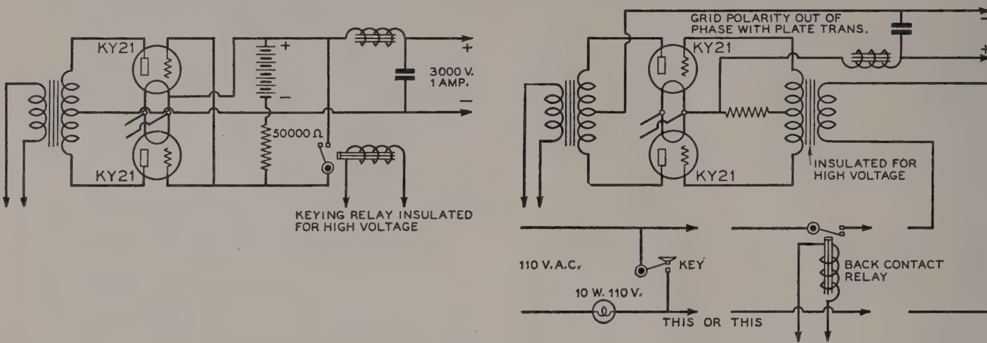


Figure 41.

GRID-CONTROLLED RECTIFIER KEYING SYSTEMS.

Two methods of using grid-controlled rectifiers for clickless, sparkless keying.

ing possible, without clicks or tails, and the plate supply to the final amplifier can be very well filtered without adding tails to the dots.

The final amplifier must have a fixed bias supply equal to more than cut-off value, so that when the grid excitation from the buffer stage is removed, the amplifier output will drop immediately to zero, in spite of the filter condenser's being fully charged in the final amplifier circuit. The bleeder across the final plate supply filter should have a very high resistance so that the filter condenser will hold its charge between dots and dashes. This will allow a quick application of plate voltage as soon as the grid excitation, supplied by the buffer stage, is applied to the final amplifier.

The buffer plate supply is keyed; its filter circuit consists of a single 2-μfd. filter condenser, shunted by the usual heavy-duty high-current bleeder resistor. This small filter has no appreciable time lag, and will not add tails to the dots and dashes, but it does provide sufficient time lag for key click elimination. The small amount of filter will not introduce a.c. hum modulation into the output of the final amplifier, because the latter is operated in class C, under saturated grid conditions. A moderate a.c. ripple in the grid excitation will not introduce serious hum in the output circuit under this operating condition.

Grid-Controlled Rectifiers. By the incorporation of grid-controlled rectifiers in a high-voltage power supply, one can enjoy keying that has practically all the advantages of primary keying with few of the disadvantages. The only disadvantage to this type of keying as compared to primary keying is that of the small amount of additional equipment needed and the added expense of the special rectifiers.

Inasmuch as no power is required to block the grids, there is little sparking at the relay contacts. And because the keying is ahead of the power supply filter, the wave train or keying envelope is rounded enough that clicks and keying impacts are eliminated. In fact, it is important that no more filter be used than is required to give a good T 9 note, inasmuch as excessive filter will introduce lag and put tails on the keying. The optimum ratio and amounts of inductance and capacity in the filter will be determined by the load on the filter (plate voltage divided by plate current). With high plate voltage and low plate current (high impedance load) more inductance and less capacity should be used, and vice versa.

Of the large number of possible circuit combinations, three of the most practical are illustrated. The circuit shown in Figure 41 at A is perhaps the simplest and most trouble-free, but has the disadvantage of requiring bias batteries. The relay contacts handle little power, but must be insulated from ground for the high voltage.

At B is shown the simplest method not requiring batteries. If used as shown, the bias transformer must be insulated for the full plate voltage (secondary to both primary and case). Unfortunately, b.c.l. transformers were not designed to withstand 3,000 or 4,000 volts r.m.s., either between windings or to the case. The circuit shown in Figure 42 allows the use of a small broadcast-receiver type transformer for bias supply to the rectifiers. In this case the whole transformer is at the power-supply voltage above ground and it must be well insulated from metal chassis and other grounded portions of the circuit.

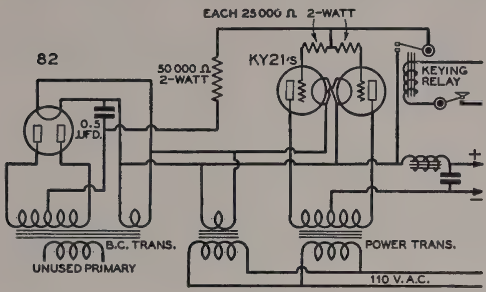


Figure 42.

ALTERNATIVE GRID-CONTROLLED RECTIFIER KEYING ARRANGEMENT.

An ordinary broadcast-receiver power transformer may be used with this circuit. The whole transformer must be well insulated from any grounded parts of the rectifier circuit.

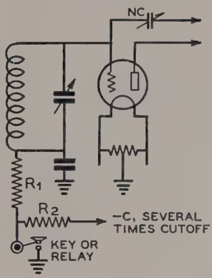


Figure 43.

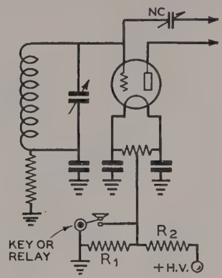


Figure 44.

ALTERNATIVE BLOCKED-GRID KEYING CIRCUITS.

Blocked Grid Keying. The negative grid bias in a medium- or low-power r.f. amplifier can easily be increased in magnitude sufficiently to reduce the amplifier output to zero. The circuits shown in Figures 43 and 44 represent two methods of such blocked grid keying.

In Figure 43, R_1 is the usual grid leak. Additional fixed bias is applied through a 100,000-ohm resistor R_2 to block the grid current and reduce the output to zero. As a general rule, a small 300- to 400-volt power supply with the positive side connected to ground can be used for the additional C-bias supply.

The circuit of Figure 44 can be applied by connecting the key across a portion of the plate supply bleeder resistance. When the key is open, the high negative bias is applied to the grid of the tube, since the filament center tap is connected to a positive point on the bleeder resistor. Resistor R_2 is the normal bleeder; an additional resistor of from one-fourth to one-half the value of R_2 is connected in the circuit for R_1. A disadvantage of this circuit is that one side of the key may be placed at a positive potential of several hundred volts above ground, with the attendant danger of shock to the operator. Blocked grid keying is not particularly effective for eliminating key clicks.

Oscillator Keying. A stable and quick-acting crystal oscillator may be keyed in the plate, cathode or screen-grid circuit for the purpose of minimizing key clicks and for break-in operation. Experience has indicated that the best method of keying a crystal oscillator is directly in the high-voltage lead.

Oscillator keying requires either fixed or cathode bias on all following r.f. stages, since the r.f. excitation is removed from all of the grid circuits. The key clicks are minimized by the presence of several tuned circuits between the antenna and crystal oscillator in a multistage transmitter. The key clicks act as sideband frequencies and are attenuated somewhat in a multistage transmitter by the resonant tuned circuits which are tuned to the carrier frequency.

If a key click filter is placed in the crystal oscillator circuit, the tone may become chirpy, and tails may be added to the ends of the transmitted characters. A practical circuit for clickless keying is illustrated in Figure 45, in which both the cathode of the crystal oscillator and the cathode of the next succeeding buffer or doubler circuit are connected through a key click filter.

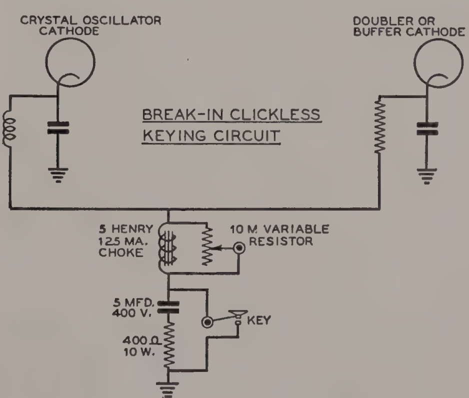


Figure 45.

BREAK-IN KEYING CIRCUIT.

This circuit arrangement may be used for break-in operation where the receiver is kept in operation on the same band as the transmitter during transmission.

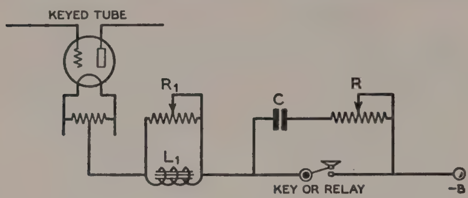


Figure 46.

CLICK FILTER CIRCUIT.

This circuit shows simple center-tap keying with an adjustable click filter to reduce interference caused by this keying method. The amount of inductance and capacity used in the filter depends upon the amount of current being keyed. Ordinarily, L_1 will be between 1 and 5 henrys, R_1 20,000 ohms, C between 0.25 and 2 μfd., and R about 2000 ohms.

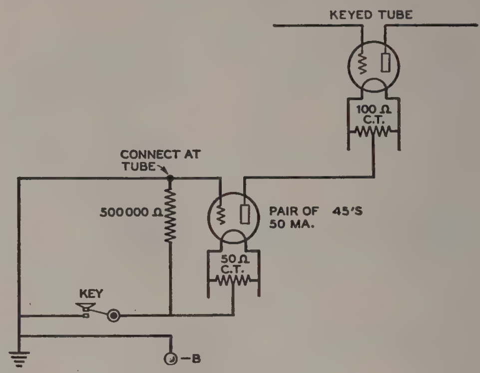


Figure 48.

VACUUM-TUBE KEYING CIRCUIT.

One of the more simple of the vacuum-tube keying circuits. Some current flows through the key in this circuit, and clicks are sometimes produced when the key is opened. Both filament transformers must be well insulated from each other and from the ground of the circuit.

Two tubes can be keyed very effectively with this type of circuit. The choke coil, shunted with a semi-variable resistor, provides a series inductance for slowing down the application of cathode current to the two tubes. The inductance of the choke coil can effectively be lowered to 1 or 2 henrys by shunting it with a semi-variable resistance so that the time lag will not be excessive. The 0.5-μfd. condenser and 400-ohm resistor are connected across the key contacts, as *close to the key* as possible, and these serve to absorb the spark at the telegraph key each time the circuit is opened. This effectively prevents a click at the end of each dot and dash. This same type of key click filter can be connected in the center-tap lead of a final amplifier or buffer-amplifier stage for the elimination of clicks.

Parasitics with Oscillator Keying. When keying in the crystal stage, or, for that matter, any stage ahead of the final amplifier, the stages following the keyed one must be absolutely stable so that parasitic or output-frequency oscillation will not occur when the excitation is rising on the beginning of each keying impulse. This type of oscillation gives rise to extremely offensive key clicks which cannot be eliminated by any type of click filter; in fact, a filter designed to slow up the rate at which signal comes to full strength may only make them worse.

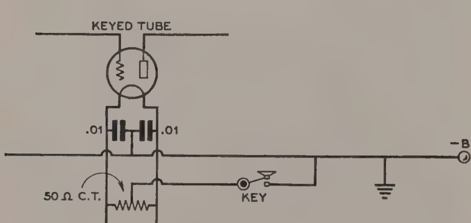


Figure 47.

ORDINARY CENTER-TAP KEYING.

The center tap of the filament transformer must not be grounded, and this transformer must not be used to supply filament voltage to any other stages.

Center-Tap Keying. The lead from the center-tap connection to the filament of an r.f. amplifier tube can be opened and closed for keying a circuit. This opens the B-minus circuit, and at the same time opens the grid-bias return lead. For this reason, the grid circuit is blocked at the same time that the plate circuit is opened, so that excessive sparking does not occur at the key contacts. Unfortunately, this method of keying applies the power too suddenly to the tube, producing a serious key click in the output circuit, which generally is coupled to the antenna. This click often can be eliminated with the key click eliminator shown in Figures 45 and 46.

Vacuum Tube Keying. Simple center-tap keying as shown in Figure 47 never should be used, because this circuit produces extremely bad key clicks. The key click filter in Figure 46 always can be connected into the center-tap lead as an external unit. A more effective key click filter for the center-tap lead is made possible through the use of vacuum tubes. A simple vacuum tube keying circuit is shown in Figure 48.

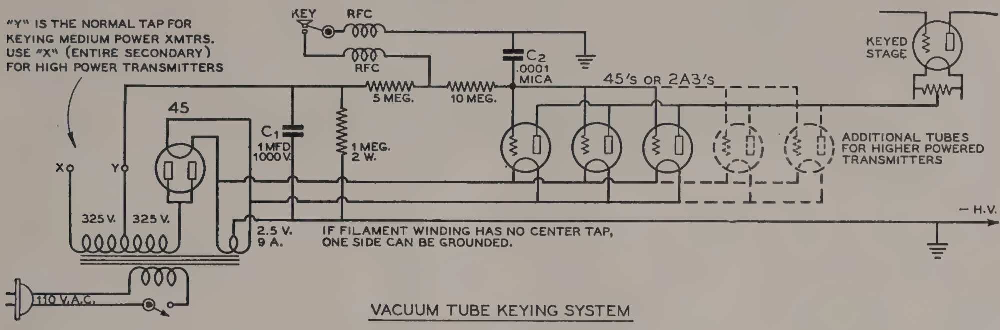


Figure 49.
VACUUM-TUBE KEYING SYSTEM.

The keying tube is connected in series with the center-tap lead of the final r.f. amplifier. The grid of the keying tube is short-circuited to the filament when the key is closed, and the keying tube then acts as a low resistance in the center-tap lead. When the key is opened, the grid of the keying tube tends to block itself and the plate-to-filament resistance of the tube increases to a high value, which reduces the output of the r.f. amplifier approximately to zero. A more effective vacuum tube keying system is shown in Figure 49.

In this system, the grids of the keying tubes are biased to a high negative potential when the key is open, and to zero potential when the key is closed. The fixed bias supply to the keying tubes provides very effective keying operation. The degree of time lag (key click elimination) can be adjusted to suit the individual operator, by varying both the capacity of the condenser which is shunted from grid to filament, and the values of the two high resistances in series with the grid and power supply leads. R.f. chokes can be connected in series with the key directly at the key terminals, to prevent the minute spark at the key contacts from causing interference in nearby broadcast receivers. These r.f. chokes are of the conventional b.c. type. There is no danger of shock to the operator when this keying circuit is used.

The small power supply for this keying circuit requires very little filter and can be of the half-wave rectifier type with a '45 tube as the rectifier. The negative voltage from this power supply needs to be sufficient only to provide cutoff bias to the type '45 keying tubes; potentials of from 100 to 300 volts are needed for this purpose. Approximately 50 ma. of plate current in the final amplifier should be allowed per type '45 keying tube. If the final amplifier draws 150 ma., for example, three type '45 keying tubes in parallel will be required.

One disadvantage of vacuum tube keying circuits is a plate supply potential loss of approximately 100 volts which is consumed by the keying tubes. The plate supply therefore should be designed to give an output of 100 volts more than ordinarily is needed for the r.f. amplifier. This loss of plate voltage is encountered because the plate-to-filament resistance of the type '45 tubes, at 50 ma. of current and zero grid potential, is approximately 2000 ohms.

Vacuum-tube keying is applicable to high-speed commercial transmitters, as well as for amateur use.

CHAPTER EIGHT

Radiotelephony Theory

RADIOTELEPHONY is the transmission by radio of audio waveforms which contain the desired intelligence of the communication. A radiotelephone transmitter differs but little, fundamentally, from a c.w. telegraph transmitter except for the audio-frequency system which is required to modulate the radiophone transmitter. Both require a frequency determining stage, a series of frequency multipliers and power amplifiers to bring the output up to the desired frequency and amount of power, and a high-voltage direct current power supply for the various stages. This chapter will deal with the theory of various modulation systems and with the theory and design of the audio channel required in the transmitter which is to be used for radiotelephony.

Modulation

Any type of continuous wave transmitter puts out a steady flow of radio-frequency energy which is varied in accordance with the intelligence which it is desired to transmit. This steady flow of r.f. energy is called the carrier or the carrier wave of the transmitter. If the outgoing carrier wave is broken up into short and somewhat longer pulses to conform with the international Morse code, the transmitter is said to be a c.w. radio telegraph transmitter. However, if either the amplitude or the frequency of the outgoing carrier wave is varied in accordance with a voice or music waveform, the source of the signal is said to be a radiotelephone transmitter. Any variation in the output of a transmitter, either c.w. or phone, is called modulation. However, the term is much more generally applied to modulation as applied to a radiophone transmitter.

Frequency Modulation, F.M. Although frequency modulation has, until recently, been thought of only as the undesirable result of amplitude modulation of the plate voltage of a self-controlled oscillator, recent developments have shown that pure frequency modulation has many advantages for certain types of work over amplitude modulation. Chief among these advantages is the improved transmission fidelity attainable through an increased signal-to-noise ratio.

In the modern system of frequency modulation, the *frequency* of the transmitter is modulated within specified limits in accordance with the voice or music to be transmitted. The amplitude of the carrier with and without modulation remains constant. A receiver which is sensitive only to variations in the frequency of the incoming carrier, and which discriminates to a large extent against variations in amplitude, is used to receive these frequency modulated signals. Since static crashes and man-made interference cause a much larger *effective* change in the amplitude of an incoming carrier than in its frequency, this system of communication gives very high quality reception with an almost total absence of noise.

Since frequency modulation has recently become of considerable importance in the field of radio communication, a complete chapter has been devoted to the theory and practice of this new subject. The reader is, therefore, referred to Chapter 9.

Amplitude Modulation, A.M. The system of modulation most widely employed at the present time for voice, music, television, facsimile, and c.w. transmission utilizes variation in the *amplitude* of the outgoing carrier in accordance with the intelligence to be transmitted. The most simple way of obtaining amplitude modulation is simply the shutting on and off of the transmitted carrier by means of a telegraph key or analogous device as used in c.w. telegraph transmission. Such keying systems are discussed in the chapter on *Transmitter Theory*. Systems of modulating the amplitude of a carrier in accordance with voice, music, or similar types of complicated waveforms are many and varied, and will be discussed later on in this chapter.

Sidebands. When a carrier wave is modulated by an audio frequency tone (varied in amplitude at an audio frequency), a result of the process of modulation is the produc-

tion of additional frequencies in the output of the transmitter which are equal to the *sum* of the carrier and the modulation frequency, and the *difference* between the two frequencies. For example, if the carrier frequency is 14,200 kc. and it is being modulated by a frequency of 2 kc. (2000 cycles) there will be two sidebands formed, one on either side of the carrier frequency. One will be equal to the sum of the two frequencies, 14,202 k.c., and the other will be equal to their difference, 14,198 kc. The frequency of the sidebands is independent of the amount of modulation of the carrier; it is determined only by the frequency of the modulating tone. This assumes, of course, that the maximum modulation capability of the transmitter is not being exceeded.

If the signal modulating the carrier consists of a number of different frequencies, as would be the case with voice or music modulation, sidebands will be formed by each of the modulating frequencies. The signal radiated by the transmitter will occupy a *band* of frequencies including the carrier and the highest modulation frequency on either side of the carrier. For example, if the highest modulation frequency is 5000 cycles, the signal emitted by the transmitter will occupy a band from 5000 cycles above to 5000 cycles below the carrier frequency. Thus the total band taken up by a carrier with 5000-cycle modulation will be *twice* the modulation frequency in width, or 10,000 cycles wide. This is true of any type of modulating waveform; the band taken up by the signal from the transmitter will be twice as wide as the highest modulation frequency.

Frequencies up to at *least* 2,500 cycles are required for good speech intelligibility, and frequencies as high as 5,000 or 6,000 cycles are required for good music fidelity.

When audio frequencies as high as 5,000 cycles are to be transmitted, the radio-frequency channel would have to be 10,000 cycles (10 kilocycles) in width, since both the upper and lower side band frequencies are generated in the modulated r.f. stage.

Since the bands of frequencies available to amateurs for radiophone transmission are limited, and since the band of frequencies taken up by a transmitter which is being modulated by needless high frequencies is quite wide, it would be advisable if all amateurs and fixed services would limit the maximum frequencies which their speech amplifiers would pass to about 3000 cycles. The passage of this frequency as maximum will give good intelligibility with a fair amount of crispness to the quality, and still the modulated carrier of the transmitter would occupy a maximum band only 6 kilocycles in width.

Mechanics of Modulation. A c.w. or unmodulated carrier wave is represented in A, Figure 1. An audio-frequency wave is represented by curve B. When this audio-frequency wave (B) is applied to the modulated stage, the resultant wave may be represented as in C and D. The *average amplitude* of the carrier wave remains constant because the decrease in amplitude is the same as the increase (up to 100 per cent). In C, Figure 1, the carrier wave is shown to be approximately 50 per cent modulated, and D shows a 100 per cent modulated wave.

In order to obtain 50 per cent modulation in a plate-modulated system, only one-fourth as much audio-frequency power is required as for 100 per cent modulation. However, the audio signal which is received at a distant point after being *demodulated* (detected) is in proportion to the percentage of modulation of the transmitter. If the peaks of modulation are reduced from 100 per cent down to

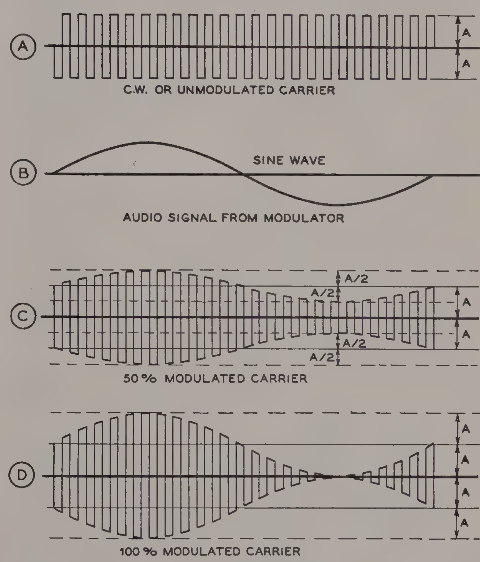


Figure 1.
MODULATION OF A CARRIER WAVE.

The top drawing (A) represents a continuous carrier wave; (B) shows the audio signal output from the modulator. (C) shows the audio signal impressed upon the carrier to the extent of 50 per cent modulation, and (D) shows the carrier with 100 per cent modulation. The carrier wave proper is also a sine wave but has been drawn in this manner to simplify the representation.

50 per cent, the result is a decrease in range of the transmitter.

The *average* amplitude of the carrier frequency wave is constant in most systems, but the *instantaneous power output* varies from approximately zero to four times that of the power of the unmodulated carrier wave. In a sinusoidally modulated wave, the antenna current increases approximately 22 per cent for 100 per cent modulation with a pure tone input; the r.f. meter in the antenna circuit indicates this increase in antenna current. The *average power* of the r.f. wave increases 50 per cent for 100 per cent modulation.

This indicates that in a plate-modulated radiotelephone transmitter, the audio-frequency channel must supply this additional 50 per cent increase in average power. If the power input to the modulated stage is 100 watts, for example, this *average power* will increase to 150 watts at 100 per cent modulation, and this additional 50 watts of power must be supplied by the *modulator* when plate modulation is used. The actual antenna power is a constant percentage of the total value of input power.

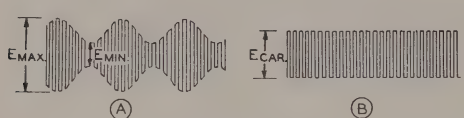


Figure 2.

MODULATED CARRIER DIAGRAM TO REPRESENT MODULATION PERCENTAGE.

Percentage of Modulation. The amount by which a carrier is being modulated may be expressed either as a modulation factor, varying from zero to 1.0 at maximum modulation, or as a percentage. The percentage of modulation is equal to 100 times the modulation factor. Figure 2A shows a carrier wave modulated by a sine-wave audio tone. A picture such as this might be seen on the screen of a cathode-ray oscilloscope with saw-tooth sweep on the horizontal plates and the modulated carrier impressed on the vertical plates. The same carrier without modulation would appear on the oscilloscope screen as Figure 2B.

The percentage of modulation of the positive peaks and the percentage of modulation of the negative peaks can be determined from two oscilloscope pictures such as shown.

The modulation factor of the positive peaks may be determined by the formula:

$$M = \frac{E_{max} - E_{car}}{E_{car}}$$

The factor for negative peaks may be determined from this formula:

$$M = -\frac{E_{min} - E_{car}}{E_{car}}$$

In the two above formulas E_{max} is the maximum carrier amplitude with modulation and E_{min} is the minimum amplitude; E_{car} is the steady-state amplitude of the carrier without modulation. Since the deflection of the spot on a cathode-ray tube is linear with respect to voltage, the relative voltages of these various amplitudes may be determined by measuring the deflections, as viewed on the screen, with a rule calibrated in inches or centimeters. The percentage of modulation of the carrier may be had by multiplying the modulation factor thus obtained by 100.

Modulation Capability. The modulation capability of a transmitter is the maximum percentage to which that transmitter may be modulated before spurious sidebands are generated in the output or before the distortion of the modulating waveform becomes objectionable. The highest modulation capability which any transmitter may have is 100 per cent modulation on the negative peaks. The maximum permissible modulation of many transmitters is less than 100 per cent. The modulation capability of a transmitter may be limited by flat tubes or by tubes with insufficient filament emission, by insufficient excitation or grid bias to a plate modulated stage, too light loading of any type of amplifier carrying modulated r.f., insufficient power output capability in the modulator, or too much excitation to a grid-modulated stage or a class B linear amplifier. In any case, the FCC regulations specify that no transmitter be modulated in excess of its modulation capability. Hence, it is desirable to make the modulation capability of a transmitter as near as possible to 100 per cent so that the carrier power may be used most efficiently.

Extended Positive Peak Modulation. The manner in which the human voice is produced by the vocal cords gives rise to a certain dissymmetry in the waveform of voice sounds when they are picked up by a good-quality microphone. This is especially pronounced in the male voice, and more so on certain voiced sounds than on others. The result of this dissymmetry in the waveform is that the voltage peaks on one side of the average value of the wave will be considerably greater, often two or three times as great, as the voltage excursions on the other side of the zero axis. The average value of energy on both sides of the wave is, of course, the same.

The net result of this dissymmetry in the male voice waveform is that especial care must be taken when modulating a transmitter if maximum sideband energy is to be obtained without distortion due to negative peak clipping or to exceeding the maximum upward modulation capability of the transmitter. The simplest solution to the problem is to pole the phase of the exciting waveform so that the large excursions in voltage are in the direction of positive modulation. This will allow a considerably higher average modulation percentage to be obtained before negative peak clipping, with its attendant serious splatter, is obtained—provided the class C amplifier is capable of *positive* modulation percentages in excess of 100 per cent. From this condition the name Extended Positive Peak Voice modulation is taken.

The most satisfactory way of determining the proper phase for the modulating waveform is to look at the modulated waveform on a cathode-ray oscilloscope, and then to reverse the polarity of the audio somewhere in the speech system while a prolonged speech sound such as " . . . errrrr" is being voiced. The polarity may be changed by reversing the microphone leads, the leads of any low-impedance line between the speech amplifier and the transmitter, or by reversing the leads to any transformer in the audio system. In either one polarity or the other the "fingers" of modulating voltage will be seen to extend in the proper upward direction.

Systems of Amplitude Modulation

There are many different systems and methods for amplitude modulating a carrier, but they may all be grouped under two general classifications: *variable efficiency* systems, in which the average input to the stage remains constant with and without modulation and the variations in the efficiency of the stage in accordance with the modulating voltage accomplish the modulation; and *constant efficiency* systems in which the input to the stage is varied by one means or another to accomplish the modulation. The various systems under each classification have individual characteristics which make certain ones best suited to particular applications.

Variable Efficiency Modulation. Since the average input remains constant in a stage employing variable efficiency modulation, and since the average power output of the stage increases with modulation, the limiting factor in such an amplifier is the plate dissipation of the tubes in the stage when they are in the unmodulated condition. Hence, for the best relation between tube cost and power output, the tubes employed should have as high a plate dissipation rating per dollar as possible.

The plate efficiency in such an amplifier is doubled when going from the unmodulated condition to the peak of the modulation cycle. Hence, the unmodulated efficiency of such an amplifier must always be less than 45 per cent, since the maximum peak efficiency obtainable in a conventional amplifier is in the vicinity of 90 per cent. Since the peak efficiency in certain types of amplifiers will be as low as 60 per cent, the unmodulated efficiency in such amplifiers will be in the vicinity of 30 per cent.

The various common systems of efficiency modulation are: grid-leak modulation, class BC grid modulation, class C grid modulation, screen-grid modulation, suppressor-grid modulation, and (a special case) cathode modulation. The class B linear amplifier also falls in this classification. Each of these various systems will be described individually.

Grid Leak Modulation. The several popular forms of grid modulation operate on the same general principle, but under somewhat different conditions. In all systems, the audio-frequency power is impressed upon the grid circuit, and the r.f. amplifier operates in a modified class C arrangement.

The simplest system employs a vacuum tube as a variable grid leak in a class C r.f. amplifier with a very small order of excitation. The modulator tube is driven by the speech amplifier, and its plate impedance varies in accordance with the speech input. The modulator tube receives its plate current from the

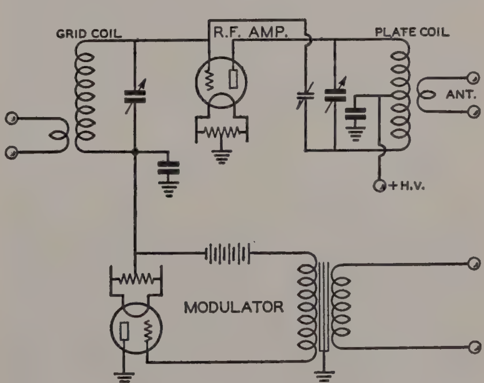


Figure 3.

SIMPLE GRID-LEAK MODULATION.

A modulation system in which the modulator tube acts as a variable grid-leak in series with the grid return of the modulated stage.

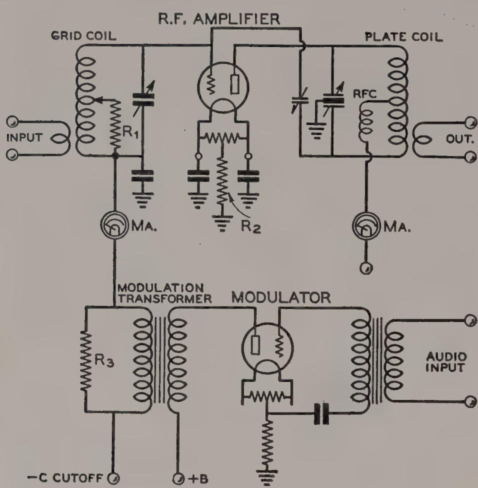


Figure 4.
CLASS BC GRID MODULATION.

A grid modulation system using an unbypassed cathode resistor in series with the return of the modulated stage to give higher plate circuit efficiency with reduced distortion due to the degenerative effect of the resistor, but requiring greater r.f. and audio drive to the stage.

rectified grid current of the r.f. amplifier. The grid bias of the modulator is adjusted to the point which gives best voice quality, and the r.f. excitation must be similarly adjusted for the same purpose. This system, shown in Figure 3, does not give distortionless modulation and is critical in adjustment. The arrangement, however, has been employed quite commonly in low-powered transmitters abroad, and is quite suitable for use in low-powered portable transmitters where it is not desired to carry heavy and bulky modulation equipment.

Class BC Grid Modulation. Figure 4 illustrates the class BC system of grid modulation. This is a system of grid modulation which can be adjusted to give exceptionally good voice quality due to the degeneration in the cathode circuit of the modulated stage.

The r.f. amplifier is operated with fixed bias equal to cutoff. This bias is supplied either from batteries or from a bias pack. Additional bias is obtained from a cathode resistor R_2 in the modulated stage. This resistor should be by-passed for r.f., but not for audio frequencies, by means of filament by-pass condensers no higher in value than .005 μfd.

When an audio voltage is applied from the modulator, it is amplified in the r.f. tube, and degenerative feedback occurs across resistor R_2. For this reason, the audio power requirements are somewhat greater than for other grid-modulated systems. This degenerative effect, however, produces a very linear modulation characteristic. The d.c. plate current which flows through R_2 should provide an additional bias equal to at least half the theoretical cutoff bias. A higher value of R_2 will result in higher plate efficiency, but at a sacrifice in power output, which can be brought up by using higher plate voltage.

The r.f. grid excitation is adjusted to the point where grid current just starts to flow. Excess r.f. grid excitation can be absorbed by resistor R_1 (Figure 4) connected across the grid circuit; this resistor also stabilizes the operation of the circuit and improves the audio quality.

Grid excitation can be conveniently controlled by means of a link-coupling adjustment. The antenna loading is greater than that required for plate modulation or c.w. operation. This coupling should be increased to a point somewhat beyond that at which maximum antenna or r.f. feeder current occurs for given excitation. The plate efficiency will be between 35 per cent and 40 per cent in a well-designed class BC amplifier.

The circuit constants can be calculated from the group of formulas given here:

(1) E_b = d.c. plate supply voltage, in volts.
(2) $W_{plate\ loss}$ = rated plate dissipation of the tube, in watts.
(3) μ = amplification factor of the tube.
(4) W_{input} = d.c. plate input power, in watts.
(5) W_{output} = r.f. unmodulated carrier output, in watts.
(6) I_p = d.c. plate current, amperes.
(7) E_{cco} = d.c. battery bias equal to theoretical cutoff bias (one-half total bias).
(8) R_k = cathode bias resistance, in ohms.
(9) $W_{input} = 1.66\ W_{plate\ loss}$
(10) $W_{output} = .66\ W_{plate\ loss}$

$$(11)\quad I_p = \frac{1.66 W_{plate\ loss}\quad (1+\mu)}{\mu E_b}$$

$$(12)\quad E_{cco} = \frac{E_b}{1+\mu}$$

$$(13)\quad R_k = \frac{E^2_b \mu}{1.66\ W_{plate\ loss}\quad (1+\mu)^2}$$

The class BC amplifier shown for grid modulation can be operated as a linear r.f. amplifier at about 35 to 40 per cent plate efficiency, which is somewhat better than the efficiency

obtainable from a conventional linear amplifier (30 to 33 per cent). However, the stage will not be a true linear amplifier since it will give an effective gain in the modulation percentage of the signal it passes. Slightly better efficiency, with less audio driving power and fewer components, may be obtained with class C grid modulation. However, if the quality of the latter system is to be improved to compare with class BC grid modulation, a separate degenerative feedback circuit external to the amplifier should be employed.

Class C Grid Modulation. The most popular and probably the most satisfactory system of grid modulation is that which is commonly called class C grid bias modulation. Figure 5 illustrates the simple circuit employed; the amplifier can, of course, be either push-pull, as indicated, or single ended as shown for the other types of grid-modulated amplifiers. The most important difference between this system and the class BC arrangement is in the amount of grid bias employed. The bias on the class BC amplifier usually runs in the vicinity of $1\frac{1}{2}$ to $2\frac{1}{2}$ times cutoff; with the class C grid bias modulation system, the bias may run as high as 8 times cutoff under the carrier conditions.

Class C grid modulation requires high plate voltage on the modulated stage, if maximum output is desired. The plate voltage is normally run about 50 per cent higher than for maximum output with plate modulation.

The driving power required for operation of a grid-modulated amplifier under these conditions is somewhat more than is required for operation at lower bias and plate voltage, but the increased power output and improved fidelity obtainable overbalances the additional excitation requirement. Actually, almost half as much excitation is required as would be needed if the same stage were to be operated as a class C plate-modulated amplifier. The resistor R across the grid tank of the stage is to serve as "swamping" to stabilize the regulation of the r.f. driving voltage. From 10 to 50 per cent of the output of the driving stage should be dissipated in this swamping resistor. If a reasonable amount of reserve excitation power is available and if a comparatively high-C grid tank is used on the grid-modulated stage, no swamping resistor will be required when the bias is at least 4 times cutoff.

A comparatively small amount of audio power will be required to modulate the amplifier stage 100 per cent. An audio amplifier having 10 watts output will be sufficient to modulate an amplifier with one kilowatt input. Proportionately smaller amounts of audio will be required for lower powered stages. However, the audio amplifier that is being used as the grid modulator should, in any case, either

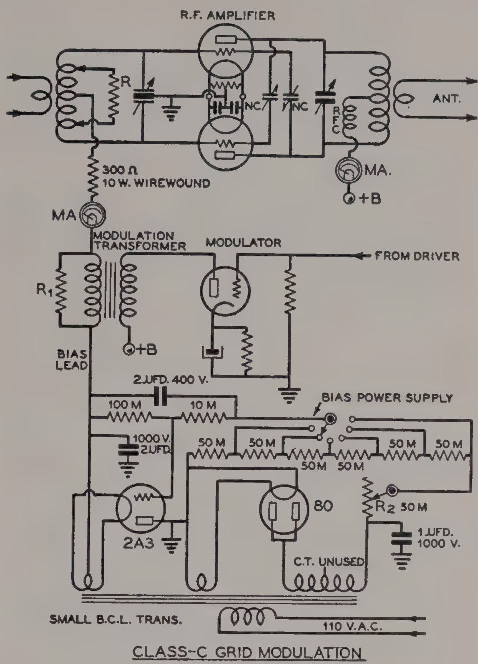


Figure 5.

employ low plate resistance tubes such as 2A3's, or else employ degenerative feedback from the output stage to one of the preceding stages of the speech amplifier. This provision of low-plate-resistance output tubes in the grid modulator is to insure good regulation in the audio driver for the grid modulated stage. Good regulation of both the audio and the r.f. drivers of a grid modulated stage is quite important if distortion-free modulation approaching 100 per cent is desired.

With the normal amount of comparatively tight antenna coupling to the modulated stage, a non-modulated carrier efficiency of 40 to 45 per cent can be obtained with substantially distortion-free modulation up to practically 100 per cent. If the antenna coupling is decreased slightly from the condition just described, and the excitation is increased to the point where the amplifier draws the same input, carrier efficiencies of 50 per cent are obtainable with tolerable distortion at 95 per cent modulation.

Tuning the Grid-Bias Modulated Stage. It will be noticed, by reference to Figure 5, that a special type of bias supply for the grid-modulated stage has been incorporated as a part of the schematic of the stage. This was done purposely to make it more clear that a special type of high-voltage bias supply is required for best operation of such an amplifier.

The arrangement shown has the advantage that the supply has very good regulation up to about 75 ma. of grid current (the maximum capability of a single 2A3) and that the voltage may be varied from nearly zero to about 700 volts; also, this particular supply may be constructed quite inexpensively.

The most satisfactory procedure for tuning a stage for grid-bias modulation of the class C type is as follows. The amplifier should first be neutralized, and any possible tendency toward parasitics under any condition of operation should be eliminated. Then a reasonable amount of antenna coupling should be made to the plate circuit, the grid bias should be run up to the maximum available value, and the plate voltage and excitation should be applied. The grid bias voltage should then be reduced until the amplifier draws the approximate amount of plate current it is desired to run, and modulation is then applied. If the plate current kicks up when a constant tone is applied, the grid bias should be reduced; if the plate meter kicks down, increase the grid bias.

When the amount of bias voltage has been found (by adjusting the fine control, R_2, on the bias supply) where the plate meter remains constant with modulation, it is more than probable that the stage will be drawing either too much or too little input. The antenna coupling should then be either increased or decreased (depending on whether the input was too little or too much, respectively) until the input is more nearly the correct value. The bias should then be readjusted until the plate meter remains constant with modulation as before. By slight jockeying back and forth of antenna coupling and grid bias, a point can be reached where the tubes are running at rated plate dissipation, and where the plate milliammeter on the modulated stage remains substantially constant with modulation.

The linearity of the stage should then be checked by any of the conventional methods; the trapezoidal pattern method employing a cathode-ray oscilloscope is probably the most satisfactory. The check with the trapezoidal pattern will allow the determination of the proper amount of gain to employ on the speech amplifier. Incidentally, too much audio power on the grid of the modulated stage should not be used in the tuning-up process, as the plate meter will kick quite erratically and it will be impossible to make a satisfactory adjustment.

Coupling Transformers for Grid Modulation. The transformer for coupling a single-ended modulator tube such as a 45 or a 2A3 into the grid circuit of the r.f. tube should preferably have a ratio of 1-to-1 or $1\frac{1}{2}$-to-1 step-down. Class AB output transformers designed for operation from 2A3's or 42 triodes into a 5000-ohm load are suitable for push-pull modulators. The shunt resistance R_1 across the secondary of the modulation transformer should be of some value between 7500 and 10,000 ohms, and should be rated at about 3 watts for the single-ended modulator and 10 watts for the push-pull class AB modulator.

Tubes for Grid Modulation. Medium-μ and high-μ triodes are most satisfactory as grid-bias modulated amplifiers. Low-μ tubes can be employed, but the amount of grid bias voltage that is required by them for high efficiency class C grid modulation is almost prohibitive unless the plate voltage is comparatively low.

Screen-grid tubes and beam tetrodes can be grid-bias modulated quite satisfactorily. The efficiency will be somewhat lower than can be obtained with triodes, but less plate voltage and considerably less excitation power is required. A very satisfactory medium-power control grid modulated phone of great compactness can be built using one of the new, high-power-gain beam tubes.

Screen-Grid Modulation. Modulation can be accomplished by varying the screen-grid voltage at an audio-frequency rate in an r.f. screen-grid tube. The screen-grid voltage must be reduced to between one-half and one-fourth the value of that used for c.w. operation. The r.f. output is correspondingly reduced and the tube then operates as an efficiency-modulated device, somewhat similar to ordinary grid modulation.

The degree of modulation is limited to approximately 90 per cent when the screen-grid of a single stage is modulated. When two cascade stages are modulated, a level of 100 per cent can be reached, with good quality. The r.f. excitation and screen-grid voltages must be carefully adjusted in order to secure satisfactory results. The r.f. excitation to the grid of the final amplifier must be so low that this tube acts somewhat like a class B linear stage.

Suppressor Modulation. Still another form of efficiency modulation can be obtained by applying audio voltage to the suppressor-grid of a pentode tube which is operated class C. A change in bias voltage on the suppressor-grid will change the r.f. output of a pentode tube, and the application of audio voltage then provides a very simple method of obtaining modulation.

The suppressor-grid is biased negatively to a point which reduces the plate efficiency to somewhat less than 35 per cent. The peak efficiency at the time of complete modulation must reach twice this value. It is difficult to obtain 100 per cent modulation, though 90 per cent to 95 per cent can easily be obtained and with good linearity.

The same modulator design problems apply to the suppressor-modulated transmitter as do to a grid-modulated amplifier. The control grid in the suppressor-modulated stage is driven to about the same degree as for c.w. or plate modulation. The r.f. excitation adjustment is not critical, but the excitation should be ample to allow distortionless modulation in this stage.

The quartz crystal should not be placed directly in the grid circuit of any suppressor-modulated stage, because of a tendency for frequency modulation and because of poor quality due to insufficient r.f. grid excitation.

A medium powered suppressor-modulated amplifier is shown in Figure 6. An 804 is used as the amplifier and will supply about 20 watts of carrier. An 803 may be substituted for the 804 to increase the carrier output to about 50 or 60 watts. Either tube may be excited to full output by a 6L6 operating either as a frequency multiplier or as a crystal oscillator. A type 45 or a 42 will serve as modulator for either tube.

It is possible to operate a suppressor-modulated amplifier stage as a doubler. The efficiency suffers somewhat but the voice quality will be found to be satisfactory.

Cathode Modulation

Within the last several years a combination modulation system called *cathode modulation* has come to the fore. In fact, it has become a quite widely used method of amplitude modulating amateur radiophone transmitters. This is easy to understand because cathode modulation offers a workable compromise between the good efficiency but expensive modulator of high-level plate modulation, and the poor efficiency but inexpensive modulator of grid modulation. Cathode modulation consists essentially of an admixture of the two, and hence can have a portion of the advantages of each with the disadvantages of neither.

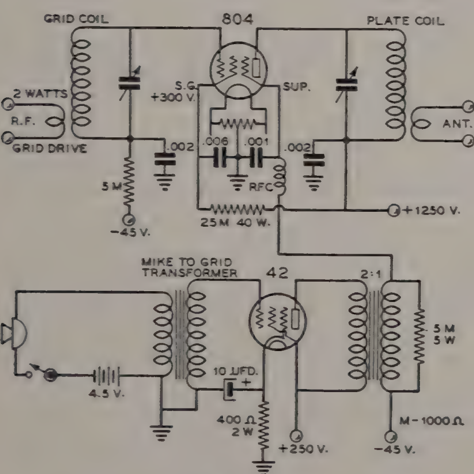


Figure 6.
LOW-POWER SUPPRESSOR MODULATED PHONE TRANSMITTER.

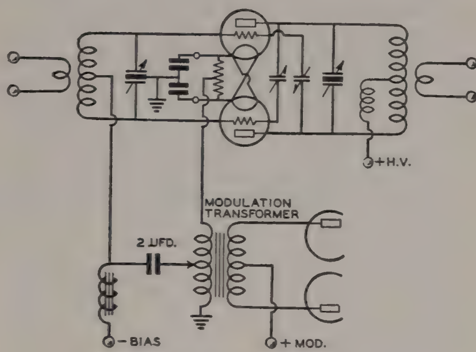


Figure 7.
CONVENTIONAL CATHODE MODULATION.

The modulation transformer in series with the cathode return of the modulated stage must match the cathode impedance of this stage. The choke in series with the grid return of the stage should have from 15 to 40 henrys inductance and should be capable of carrying the full grid current of the stage. The grid tap on the modulation transformer is varied, after the stage has been placed into operation, to give the best modulation pattern as the carrier is viewed on the screen of a cathode-ray oscilloscope.

The efficiency of the average well-designed plate-modulated transmitter is in the vicinity of 75 to 80 per cent, with a compromise perhaps at 77.5 per cent. On the other hand, the efficiency of a good grid-modulated transmitter may run from 28 to maybe 44 per cent, with the average falling toward the lower limit at about 33 per cent. Now since cathode modulation consists of simultaneous grid and plate modulation, in phase with each other, we can theoretically obtain any efficiency from about 33 to 77.5 per cent from our cathode-modulated stage, depending upon the relative percentages of grid and plate modulation.

Since the system is a compromise between the two fundamental modulation arrangements, and the relative values of the two are approximately the same, a value of efficiency approximately half way between the two would seem to be the best compromise. Experience has consistently proven this to be the case. A

compromise efficiency of about 56.5 per cent, roughly half way between the two limits, has proven to be optimum. Calculation has shown that this value of efficiency will be obtained from the normal cathode-modulated amplifier when the audio frequency modulating power is approximately 20 per cent of the d.c. input to the cathode-modulated stage.

The Cathode-Modulation Operating Curves. Figure 8 shows a set of operating curves for cathode-modulated r.f. amplifier stages. The chart is a plot of the percentage of plate modulation (*m*) against plate circuit efficiency, audio power required, plate input wattage in per cent of the plate-modulated class C rating, and output power in percentage of the class C phone output rating. These last two curves are not of as great importance to the amateur designing a new transmitter as are the curves showing the relationship between per cent plate modulation and plate circuit efficiency. Since the relation between per cent plate modulation and the audio power output of the cathode modulator expressed as a percentage of the input to the modulated stage is a linear one, the output power of the modulator (as a percentage of the modulated stage input) is one-half the per cent plate modulation.

Optimum Operating Conditions. As was mentioned before, the optimum operating condition for a normal cathode modulated amplifier is that at which the audio power output of the cathode modulator is about 20 per cent of the d.c. input to the modulated stage. Under these conditions the plate efficiency will be in the vicinity of 56.5 per cent (between 54 and 58 per cent in a practical transmitter), the permissible power input for tubes of late design will be about 65 per cent of the rated class C *modulated* input to the tube. On tubes which do not have ICAS ratings, this value of input will be considerably less than the maximum that the tube can handle. The limiting factor in an efficiency modulated amplifier of this type is, to a large extent, plate dissipation. If, under the conditions given above, the plate dissipation of the tube under carrier conditions is less than the rated value, the plate input can be increased until rated plate dissipation is reached. This, in the case of the older tubes, will allow a considerably greater power output to be obtained. The plate dissipation for any condition of operation can easily be determined by reference to Figure 9 and a little calculation. Determine the input from Figure 8, and from the efficiency value given in either Figure 8 or 9, figure the power output from the stage—subtract this from the plate input, and the result is the amount that the tube will be required to dissipate.

Cathode Impedance. The impedance of the cathode circuit of an amplifier, which is being cathode modulated, is an important consideration in the selection of the transformer which is to be used to couple the modulator to the stage. The cathode impedance of an amplifier is equal to the peak *modulating* voltage divided by the peak a.f. component of the plate current of the stage. The peak *modulating* voltage is equal to the plate voltage times *m* (the per cent plate modulation, or *twice* the percentage of audio used to modulate the amplifier).

$$\text{Hence: } Z_k = m\frac{E_p}{I_p}$$

Or, simply, the cathode impedance is equal to the per cent plate modulation (expressed as a factor, as 0.4 for 40 per cent plate modulation) times the plate voltage, divided by the plate current.

The Cathode Modulator. The modulator which is used to feed the audio into the cathode circuit of the modulated stage should have a power output of 20 per cent of the d.c. input to the stage for 40 per cent plate modulation. Although this is the recommended percentage of plate modulation, satisfactory operation may be had with other percentage values than this, provided the proper operating values are taken from the charts of Figures 8 and 9. The modulator tubes may be operated class A, class AB, or class B, but it is recommended that some form of degenerative feedback be employed around the modulator tubes when they are to be operated in any manner other than class A. This is particularly true of beam tetrodes when used as modulators; if some form of feedback is not used around them the harmonic distortion can easily be serious enough to be objectionable, since the cathode modulated stage does not present a strictly linear impedance.

The transformer which couples the modulator to the cathode circuit of the modulated amplifier should match the cathode impedance, as calculated by the formula above, and in addition should have a number of taps so that the proper amount of audio voltage will be impressed upon the grid of the stage. In most cases one of the conventional multi-match output transformers will be satisfactory for the job, the cathode lead and the ground terminal of the stage being connected to the proper taps to give the desired value of impedance. The stage is then coupled to a cathode-ray oscilloscope so that the modulated waveform is shown on the screen. As the stage is being modulated, the grid is tapped varying amounts up and down on the modulation transformer until the best waveform is obtained on the

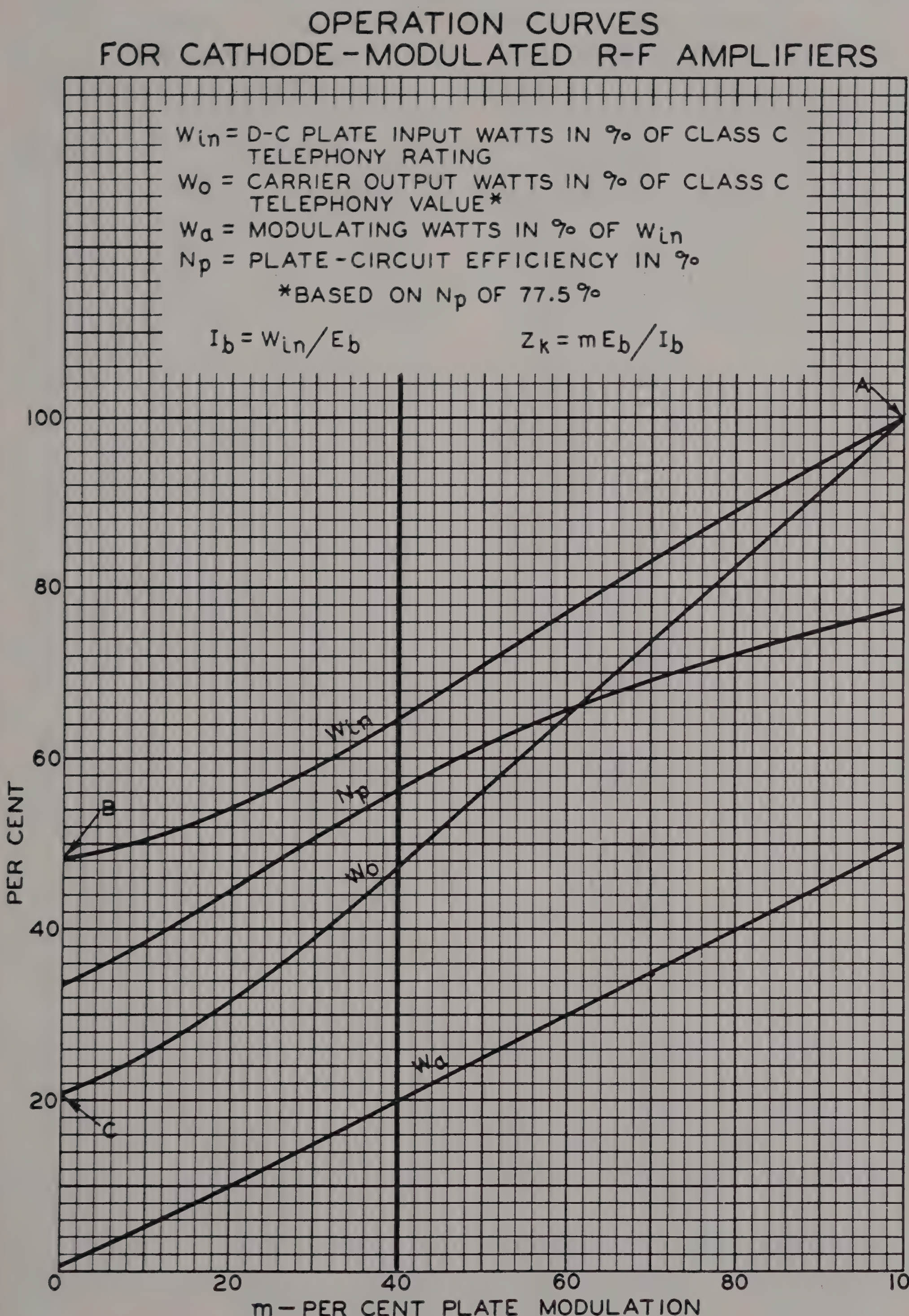


FIGURE 8.

Courtesy R C A Mfg. Co.

screen of the oscilloscope. The more closely the grid is tapped to the cathode, the less will be the amount of audio voltage upon the grid. On the other hand, if the grid return is grounded, the full cathode swing will be placed upon the grid. It will be found that low-μ tubes will require a larger percentage of the total cathode swing upon them than will tubes with higher μ factors. Hence, high-μ tubes will be tapped closer to cathode; low-μ tubes will be tapped more closely to ground.

Excitation. The r.f. driver for a cathode-modulated stage should have about the same power output capabilities as would be required to drive a c.w. amplifier to the same input as it is desired to drive the cathode-modulated stage. However, some form of excitation control should be available since the amount of excitation power has a direct bearing on the linearity of a cathode-modulated amplifier stage. If link coupling is used between the driver and the modulated stage, variation in the amount of link coupling will afford ample excitation variation.

Biasing Systems. Any of the conventional biasing arrangements which are suitable for use on a class C amplifier are also suitable for use with a cathode-modulated stage. Battery bias, grid leak bias, and power supply bias all are usable in their conventional fashion; cathode bias may be used if the bias resistor is bypassed with a high-capacity electrolytic condenser. In any case the bias voltage should be variable or adjustable so that the optimum value for distortionless modulation can be found. If grid leak or cathode bias is used, the value of the grid leak or cathode resistor should be adjustable.

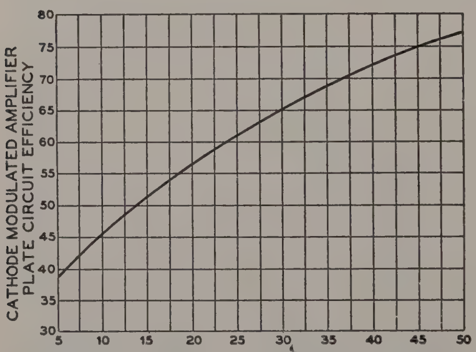


Figure 9.
EFFICIENCY OF A CATHODE MODULATED AMPLIFIER AS A FUNCTION OF THE POWER OUTPUT OF THE MODULATOR STAGE.

Parallel and Series Cathode Modulation

There are two additional methods of cathode modulation that deserve mention, since they offer certain advantages for various types of circuit applications. Both parallel and series cathode modulation, as they have been called, use a single-ended cathode modulator stage, and neither requires a modulation transformer.

Parallel Cathode Modulation. This circuit is an adaptation of the "cathode follower" type of audio amplifier in which the cathode currents of both the modulator stage and the modulated amplifier flow through a common cathode choke, with the plate of the modulator tube returned to ground. The two main advantages of this circuit arrangement are that: First, since the cathode impedance of the modulator tube is very closely the same as that of the cathode circuit of the tube to be modulated, no matching transformer is needed as a coupling impedance between them; it is only necessary to insert an ordinary choke capable of carrying the sum of their plate currents in the common cathode circuit of the two tubes. Second, the tubes in the parallel cathode modulator are operating with 100 per cent degenerative feedback; the plate is returned to the h.v. power supply and all the audio voltage output of the stage is impressed upon the grid of the tube 180° out of phase with the incoming voltage.

Degenerative Feedback. It is this inherent degenerative feedback which lowers the effective plate impedance (or cathode impedance, if you want to call it that) to such a great extent. Although the percentage of feedback will be the same in all cases (100 per cent), the value of feedback expressed in decibels will be a function of the effective amplification of the tubes in the modulator.

Due to the fact that there is such a large amount of degenerative feedback around the modulator tube, distortion within this tube will be greatly minimized. Single-ended beam tubes operated in the conventional manner have a rather large percentage of harmonic distortion: 8 to 15 per cent is not uncommon. However, with the audio energy being taken from the cathode of the tube, not only is the plate impedance lowered, but the harmonic distortion is lowered by a comparable amount.

Another thing that will be noticed by reference to the paragraph under *Degenerative Feedback* is that the voltage swing required on the grid of the modulator tube is quite high, and that it is determined almost primarily by the power output of the tube rather than by the amplification factor as it would be in a

conventional amplifier stage. In any case, a peak voltage of 200–500 volts should be figured upon to excite the grid of the parallel-cathode modulator.

Series Cathode Modulation. This is a system of cathode modulation which is ideally suited as an alternative modulating arrangement for a high-power c.w. transmitter. The modulator can be constructed quite compactly and for a minimum cost for components, since no power supply is required for it. When it is desired to change over to 'phone, it is only necessary to plug the series cathode modulator into the cathode return-circuit of the c.w. amplifier stage—the plate supply for the modulator tubes and for the speech amplifier is taken from the cathode voltage drop of the modulated amplifier across the modulator unit.

Figure 11 shows the circuit of a modulator which was used to cathode modulate a pair of 810's with 660 watts of plate input. The tubes in the modulated amplifier ran at about 50 per cent plate efficiency, giving an output of about 350 watts. In this arrangement, the final stage ran at a plate voltage of 2400 volts, with about 400 volts drop across the cathode modulator, giving a net plate to cathode voltage at the tubes of about 2000 volts. The plate current was 330 ma., the grid current was approximately 40 ma., making the total cathode current of the stage about 370 ma. The four paralleled 6L6's were able to handle this plate current without difficulty. Similar applications of the system may be made in the case of the majority of c.w. transmitters which normally run from 500 to 1000 watts input. The input will, of course, have to be reduced a substantial amount when the stage is being cathode modulated.

The Class B Linear Amplifier. The operation of the class B linear amplifier has been discussed in the chapter devoted to vacuum-tube theory, and hence will only be covered quite generally here. The linear amplifier is not well suited for use in amateur stations, since the value of unmodulated plate efficiency is quite low, varying from 30 to 39 per cent.

The grid circuit of a linear amplifier is fed modulated r.f. energy and the stage amplifies carrier and sidebands linearly. The stage is biased to cutoff with no excitation so that when excitation is applied the plate current flows in 180° pulses. This long period of plate current flow limits the theoretical peak efficiency to 78.5 per cent, the practical peak efficiency to about 65 per cent, and the average carrier efficiency to, as mentioned before, about 30 to 33 per cent for a conventional linear. The efficiency in a class BC linear, due to the greater grid bias and more narrow angle of flow, can be as high as 40 per cent.

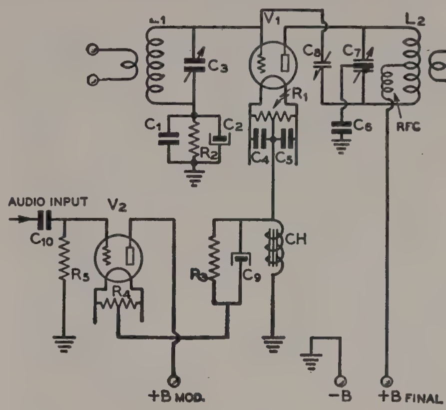


Figure 10.

PARALLEL CATHODE MODULATED R.F. STAGE.

Wiring diagram of a single-ended r.f. amplifier being modulated by a triode parallel cathode modulator. R_3, by-passed by C_9, plus the drop across the choke CH furnishes the bias for the modulator tube V_2.

- C_1—.005-μfd. mica
- C_2—8-μfd. 450-volt elect
- C_3—Conventional for band of operation
- C_4, C_5—.001-μfd. mica
- C_6—.002-μfd. 5000-volt mica
- C_7—Conventional for band of operation
- C_8—Neut. condenser for V_1
- C_9—10-μfd. 100-volt elect.
- C_{10}—.05-μfd. 400-volt tubular
- R_1—C.t. resistor or tap on fil. trans.
- R_2—10,000 to 25,000 ohms, 10 watts
- R_3—500 ohms, 10 watts
- R_4—C.t. res. or tap on fil. trans.
- R_5—100,000 ohms, 1 watt
- L_1, L_2—Coils for band operated
- RFC — Conventional for band operated
- CH—3-to-10 hy. 150-ma. filter choke
- V_1—809, T-20, or similar
- V_2—2A3 or pair 45's in parallel

The power output from a correctly operating linear amplifier will be about one half the maximum plate dissipation of the stage under the carrier conditions. The schematic of a linear amplifier is exactly the same as a conventional amplifier, whether single ended or push-pull, except that a swamping resistor is usually placed across the grid circuit of the stage. A linear amplifier generally requires from 5 to 10 per cent as much excitation power as will be obtained from its output circuit.

Input Modulation Systems

Constant efficiency variable-input modulation systems operate by virtue of the addition of external power to the modulated stage to

effect the modulation. There are two general classifications that come under this heading; those systems in which the additional power is supplied as audio frequency energy from a modulator, usually called plate modulation systems, and those systems in which the additional power to effect modulation is supplied as direct current from the plate supply.

Under the former classification comes Heising modulation (probably the oldest type of modulation to be applied to a continuous carrier), class B plate modulation, and series modulation. These types of plate modulation are by far the easiest to get into operation, and they give a very good ratio of power input to the modulated stage to power output; 65 to 80 per cent efficiency is the general rule. It is for these two important reasons that these modulation systems, particularly class B plate modulation, are at present the most popular among amateurs.

Modulation systems coming under the second classification are of comparatively recent development and have only quite recently been applied even to broadcast work. There are quite a few systems in this class but only two are really worthy of special consideration. These are: the Doherty linear amplifier, and the Terman-Woodyard high-efficiency grid modulator. Both systems operate by virtue of a carrier amplifier and a peak amplifier connected together by means of electrical quarter-wave lines. They will be described later on in this section.

Plate Modulation. Plate modulation is the application of the audio power to the *plate circuit* of an r.f. amplifier. The r.f. amplifier must be operated class C for this type of

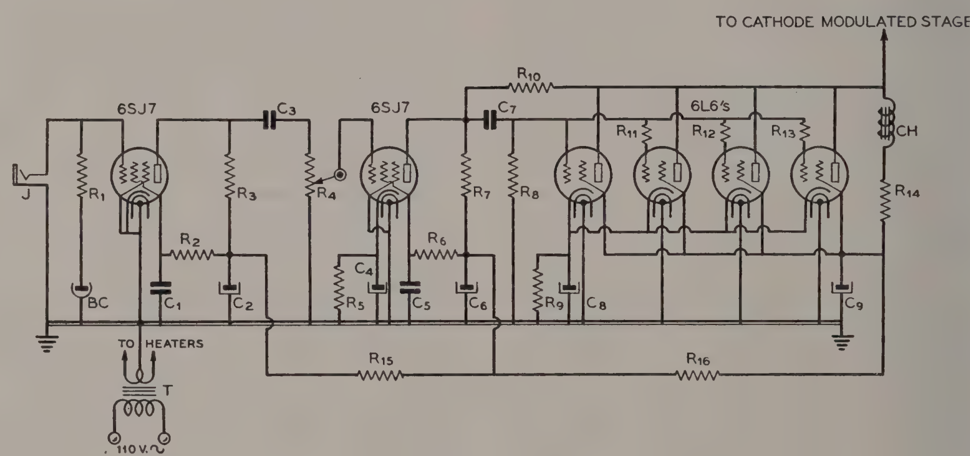


Figure 11.

SERIES CATHODE MODULATOR FOR A HIGH-POWERED R.F. AMPLIFIER.

Wiring diagram of the series cathode modulator using four paralleled 6L6's in the output. The first two stages of speech up to and including R_{10} and R_8 would remain the same for different output stages.

C_1—0.25-μfd. 600-volt tubular
C_2—8-μfd. 450-volt electrolytic
C_3—.02-μfd. 400-volt tubular
C_4—10-μfd. 25-volt tubular
C_5—0.1-μfd. 600-volt tubular
C_6—8-μfd. 450-volt electrolytic
C_7—.05-μfd. 400-volt tubular
C_8—50-μfd. 50-volt electrolytic
C_9—8-μfd. 450-volt electrolytic
R_1—1 megohm, ½ watt
R_2—500,000 ohms, ½ watt
R_3—250,000 ohms, ½ watt
R_4—500,000-ohm potentiometer
R_5—500 ohms, 1 watt
R_6—500,000 ohms, ½ watt
R_7—100,000 ohms, 1 watt
R_8—100,000 ohms, ½ watt
R_9—50 ohms, 10 watts
R_{10}—500,000 ohms, ½ watt
R_{11}, R_{12}, R_{13}—100 ohms, 1 watt
R_{14}—2500 ohms, 10 watts
R_{15}—10,000 ohms, 1 watt
R_{16}—10,000 ohms, 1 watt
T—6.3-volt 6-ampere filament transformer
CH—10-hy. 65-ma. filter choke
J—Crystal mike jack
BC—Bias cell

modulation in order to obtain a radio-frequency output which changes in exact accordance with the variation in plate voltage. *The r.f. amplifier is 100 per cent modulated when the peak a.c. voltage from the modulator is equal to the d.c. voltage applied to the r.f. tube.* The positive peaks of audio voltage increase the instantaneous plate voltage on the r.f. tube to *twice* the d.c. value, and the negative peaks reduce the voltage to zero.

The instantaneous plate *current* to the r.f. stage also varies in accordance with the modulating voltage. The peak alternating current in the output of a modulator must be equal to the d.c. plate current of the class C r.f. stage at the point of 100 per cent modulation. This combination of change in audio voltage and current can be most easily referred to in terms of *audio power in watts.*

The plate efficiency of the plate-modulated stage is constant, and the additional power radiated in the form of sidebands is supplied by the modulator.

One of the advantages of plate (or power) modulation is the ease with which proper adjustments can be made in the transmitter. There is less plate loss in the r.f. amplifier for a given value of carrier power than with other forms of modulation, because the plate efficiency is higher.

By properly matching the plate impedance of the r.f. tube to the output of the modulator, the ratio of voltage and current swing to d.c. voltage and current is automatically obtained. The modulator should have a peak voltage output equal to the average d.c. plate voltage on the modulated stage. The modulator should also have a *peak power* output equal to the d.c. plate input power to the modulated stage. The *average* power output of the modulator will depend upon the type of waveform and upon the type of modulator. If the amplifier is being Heising modulated by a class A stage, the modulator must have an average power output capability of one-half the input to the class C stage. If the modulator is a class B audio amplifier, the average power required of it may vary from one-quarter to one-half the class C input depending upon the waveform. However, the *peak* power output of any modulator must be equal to the class C input to be modulated. This subject is completely covered in the section *Speech Waveforms.*

Heising Modulation. Heising modulation is a system of plate modulation, and usually consists of a class A audio amplifier coupled to the r.f. amplifier by means of a modulation choke coil, as shown in Figure 12.

The d.c. plate voltage and plate current in the r.f. amplifier must be adjusted to a value which will cause the plate impedance to match

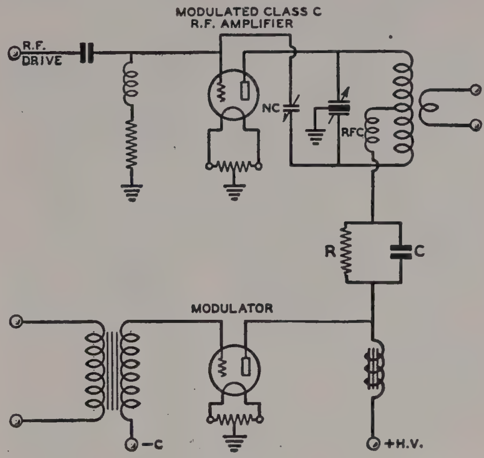


Figure 12.

HEISING PLATE MODULATION.

The resistor R in series with the lead to the class C amplifier drops the plate voltage to a value which will allow a high degree of modulation. The condenser C by-passing it can be 2 to 4 μfd. in capacity and should be able to withstand about the same amount of voltage as is placed upon the modulator for a good safety factor.

the output of the modulator, since the modulation choke gives a 1-to-1 coupling ratio. A series resistor, by-passed for audio frequencies by means of a condenser, must be connected in series with the plate of the r.f. amplifier in order to obtain modulation up to 100 per cent. The a.c. or audio output voltage of a class A amplifier does not reach a value equal to the d.c. voltage applied to the class A amplifier and, consequently, the d.c. plate voltage impressed across the r.f. tube must be reduced to a value equal to the maximum available a.c. peak voltage.

A higher degree of distortion can be tolerated in low-power emergency phone transmitters which use a pentode modulator tube for securing sufficient audio output, and thus the series resistor and by-pass condenser are usually omitted.

Class B Plate Modulation. High-level class B plate modulation is probably the most satisfactory and least expensive method of plate modulating inputs of from 50 to 1000 watts. Since most amateur phone transmitters fall into this power range, considerable discussion will be given to the various problems associated with the design problems of class B modulators. Figure 13 shows a conventional class B plate-modulated class C amplifier.

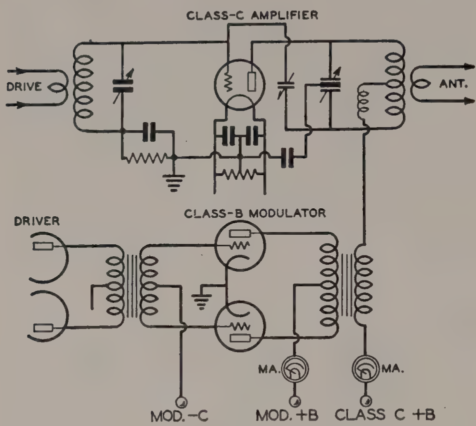


Figure 13.
CLASS B PLATE MODULATION.

The statement that the average modulator power must be one-half the class C input for 100 per cent modulation is correct only if the waveform of the modulating power is a *sine wave.* For amateur purposes, where the modulator waveform is speech, the average modulator power for 100 per cent modulation is considerably less than one-half the class C input. If a modulator is to be used *only with speech,* it seems logical to assume that its design be based upon the peculiarities of speech rather than on the characteristics of the sine wave. The difference between speech and the sine wave is so pronounced that a 100-watt class B modulator, *if properly designed for speech,* may be used to modulate fully an input of from 300 to 400 watts. The idea cannot be applied to Heising modulators (class A single ended) for reasons that will be apparent when it is recalled that such modulators run hottest when resting and that the plate operating voltage limits the peak output as well as the average output.

Power Relations in Speech Waveforms. It has been determined experimentally that the ratio of peak to average power in a speech waveform is approximately 4 to 1 as contrasted to a ratio of 2 to 1 in a sine wave. This is due to the high harmonic content of such a waveform, and to the fact that this high harmonic content manifests itself by making the wave unsymmetrical and causing sharp peaks or "fingers" of high energy content to appear. It follows from this that, for speech, the *average* modulator plate current, plate dissipation, and power output are approximately one-half the sine wave values for a given *peak* output power. In other words, a 100-watt class B modulator, if used to modulate 100 per cent with speech an input of 200 watts, delivers an *average* power of only about 50 watts and the average plate current and plate dissipation are only one-half the permissible values. In order to take full advantage of the tube ratings, the design should be altered so that the *peak* power output is increased until the average plate current or plate dissipation becomes the limiting factor.

Both peak power and average power are necessarily associated with waveform. *Peak* power is just what the name implies: the power at the peak of a wave. Peak power, although of the utmost importance in modulation, is of no great significance in a.c. power work, except insofar as the *average* power may be determined from the peak value of a known wave form. There is no time element implied in the definition of peak power; peak power may be instantaneous—and for this reason average power, which is definitely associated with time, is the important factor in plate dissipation. It is possible that the peak power of a given wave form be several times the average value; for a sine wave, the peak power is twice the average value, and for speech the peak power is approximately four times the *average* value. For 100 per cent modulation, the *peak* (instantaneous) audio power must equal the class C input, although the average power for this value of peak varies widely depending upon the modulator wave form, being 50 per cent for a sine wave and about 25 per cent for speech tones. The problem then of obtaining more speech power consists in obtaining as high a *peak* power as possible without exceeding the *average* plate dissipation or current rating of the tubes.

Since the power output varies as the square of the peak current, the most logical thing to do in order to obtain high peak power is to increase the peak current. This may be done by decreasing the class B modulator plate-to-plate load.

At this point it might be assumed that this increase in peak current is nothing more or less than a gross overload without regard for the manufacturer's ratings. However, a little reflection will show that the manufacturer's rating is given as *average* current and that the actual *peak* current (this cannot be read by a meter) varies widely with the mode of operation. An average plate current of 100 ma. in class C operation may call for a dynamic peak plate current of 1 ampere, whereas in class B service this same 100 ma. per tube represents a peak of only 315 ma. No ill effects will result if the peak plate current is increased to such a point that the average plate current with speech, as read on the plate meter, is

equal to the sine-wave value as specified by the manufacturer. With this in mind, the *peak* plate current may be safely doubled, assuming that the plate dissipation does not become the limiting factor.

Modulation Transformer Calculations. The modulation transformer is a device for matching the load impedance of the class C amplifier to the recommended load impedance of the class B or possibly class AB modulator tubes. Modulation transformers are usually designed to carry the class C plate current through their secondary windings, as shown in Figure 13. The manufacturer's ratings should be consulted to insure that the d.c. plate current being pulled through the secondary winding does not exceed the maximum rating.

The load resistance presented by the class C r.f. amplifier to the modulation transformer is calculated by dividing the d.c. plate-to-filament voltage by the plate current of the stage. For example, a pair of 75T tubes in a push-pull amplifier operating at 1200 volts and 250 milliamperes present a load impedance of 1200 divided by 0.25 amperes, or 4800 ohms.

$$Z = \frac{E}{I} = \frac{1200}{0.25} = 4800 \text{ ohms},$$

where Z is the load impedance of the class C r.f. amplifier.

The power input is 1200 times 0.25 or 300 watts.

By reference to Figure 14 we see that a pair of 809's operating at 750 volts will fully modulate an input of 300 watts to a class C amplifier with voice-waveform audio power. In other words, the peak audio output of the class B 809's, when operated into a load impedance of 4800 ohms and at a plate voltage of 750, is 300 watts. It just so happens that the recommended plate-to-plate load resistance of the 809's under these operating conditions is the same as the load presented by the class C amplifier. Hence, the modulation transformer should have a primary-to-secondary ratio of 1-to-1. The other operating conditions for the 809 modulator will be found in Figure 14. A modulation transformer rated to handle 125 watts of audio will be ample for the purpose.

Suppose we take as another example, to illustrate the method of calculation, the case of a pair of 54 Gammatrons operating at 2000 volts at 250 ma. This amplifier would present a load resistance of 2000 divided by 0.25 amperes or 8000 ohms. The plate power input would be 2000 times 0.25 amperes or 500 watts. By reference to Figure 14 we see that a pair of TZ40's at 1000 volts will put out 500 peak audio watts and hence will modulate 500 watts input with speech waveform. The plate-to-plate load for these tubes is given as 5100 ohms; hence, our problem is to match a load of 8000 ohms to the proper load resistance of the TZ40's of 5100 ohms.

A 200-to-300 watt audio transformer will be required for the job. If the taps on the transformer are given in terms of impedances it will only be necessary to connect the secondary for 8000 ohms and the primary to 5100 ohms. If it is necessary to determine the turns ratio of the transformer it can be determined in the following manner. The square root of the impedance ratio is equal to the turns ratio, hence:

$$\sqrt{\frac{8000}{5100}} = \sqrt{1.57} = 1.25$$

The transformer must have a turns ratio of 1.25 to 1, step up. The transformer must be step-up since the higher impedance is on the secondary. When the primary impedance is the higher of the two impedances, the transformer must be connected step-down.

Bass Suppression. Not only can a smaller class B modulator be used for complete modulation of a given carrier power when voice only is to be used, but an increase in the effectiveness of the modulator power can be ob-

Figure 14.
CLASS C INPUT THAT CAN BE FULLY SPEECH-MODULATED BY VARIOUS CLASS B TUBES

Class B Tubes	*Class C Power Input*	*Class B P-P Load*	*Plate Voltage*	*Average Speech Plate Current*	*Class B Bias*	*Driver Tubes*	*Average Driving Power*	*Driver Transformer Ratio Pri. to ½ Sec.*
TZ-20	*250*	*4850*	*750*	*145*	*0*	*2-2A3*	*7*	*2.6:1*
809	*300*	*4800*	*750*	*165*	*—1½*	*2-2A3*	*5*	*4.5:1*
809	*400*	*7200*	*1000*	*150*	*—8*	*2-2A3*	*5*	*4.5:1*
TZ-40	*500*	*5100*	*1000*	*200*	*—5*	*2-2A3*	*8*	*2.6:1*
TZ-40	*600*	*7400*	*1250*	*182*	*—9*	*2-2A3*	*7*	*2.8:1*
203Z	*800*	*5500*	*1250*	*250*	*0*	*4-2A3*	*15*	*2.75:1*

tained by incorporation of a simple bass suppression circuit. Most of the audio power generated in a modulator is represented by the bass frequencies. As the frequencies below 200 or 250 cycles can be greatly attenuated without noticeably affecting the speech intelligibility, it is desirable to do so for communication work. Bass suppression permits a higher percentage modulation at the voice frequencies providing intelligibility, which is equivalent to a substantial increase in power. It is not necessary to suppress the bass frequencies completely; but only to attenuate them until, as the audio gain is increased, overmodulation first occurs at the voice frequencies that afford intelligibility, rather than at the power-consuming bass frequencies.

In Figures 15 and 16 are shown two simple systems for bass suppression. They are self-explanatory and can be placed between almost any two voltage amplifier tubes in your speech channel. They will work into or out of either triodes or pentodes, but *don't use inverse feedback around the suppressor* or you'll suppress the suppression!

The bass suppressor is an old idea in the talking picture field. It is really surprising how much it cleans up the average boomy ham quality on voice. One reason the new F-type telephone handset mikes sound so good on speech is that they cut off very sharply below 200 cycles.

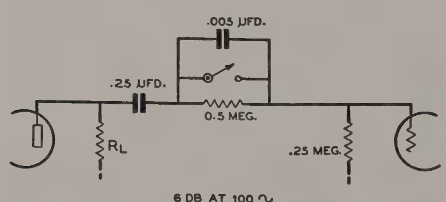
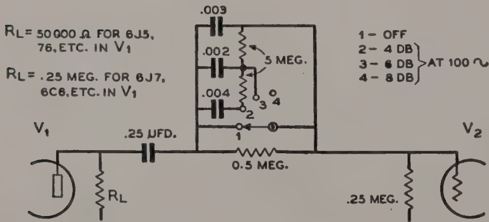


Figure 15.

Simple dialogue equalizer that can be installed in any speech amplifier.

Figure 16.

Variable bass suppression dialogue equalizer.



The bass suppressor shown in Figure 15 has a suppression of 6 db at 100 cycles while the arrangement of Figure 16 has 0, 4, 6 and 8 db suppression in the four switch positions. The 5-megohm resistors merely eliminate the loud clicks which otherwise would be heard when varying the suppression.

In both of the arrangements, the suppression starts at about 500 cycles although the good work really begins below 200 cycles. The 1000-cycle gain of an amplifier equipped with this type of bass suppression is practically unchanged with the suppressor in or out.

Plate-and-Screen Modulation. When *only* the plate of a screen-grid tube is modulated, it is impossible to obtain high percentage linear modulation, except in the case of certain beam tubes. A dynatronic action usually takes place when the instantaneous plate voltage falls below the d.c. screen voltage, and this prevents linear modulation. However, if the screen is modulated simultaneously with the plate, the instantaneous screen voltage drops in proportion to the drop in the plate voltage, and linear modulation can then be obtained. A circuit for such a system is shown in Figure 17.

The screen r.f. by-pass condenser, C_2, should not have a value greater than .01 μfd., preferably not larger than .005 μfd. It should be large enough to bypass effectively all r.f. voltage without short-circuiting high-frequency audio voltages. The plate by-pass condenser can be of any value from .002 μfd. to .005 μfd. The screen-dropping resistor, R_2, should reduce the applied high voltage to the value specified for operating the particular tube in the circuit. Condenser C_1 is seldom required, yet some tubes may require this condenser in order to keep C_2 from attenuating the high audio frequencies. Different values between .001 and .0002 μfd. should be tried for best results.

Another method is to have a third winding on the modulation transformer, through which the screen-grid is connected to a low-voltage power supply. The ratio of turns between the two output windings depends upon the type of screen-grid tube which is being modulated. The latter arrangement is more economical insofar as modulator power is concerned, because there is no waste of audio power across a screen-grid voltage-dropping resistor. However, this loss is relatively small anyway with most tubes. The special transformer is not justified except perhaps for high power.

Quite good linearity at high percentage modulation can be obtained with some of the *beam*-type transmitting tetrodes by modulating the plate voltage alone.

If the screen voltage for the beam tube is derived from a dropping resistor (*not* a divider) that is by-passed for r.f. but not a.f.,

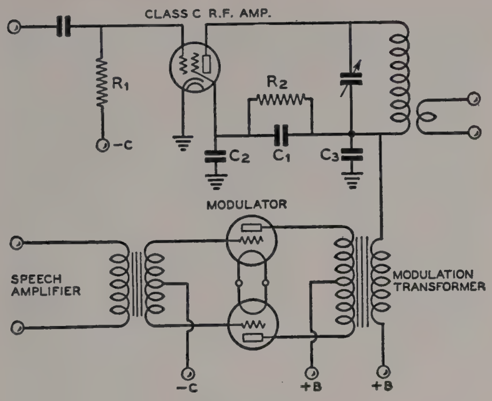


Figure 17.
PLATE AND SCREEN MODULATION OF AN R.F. STAGE.

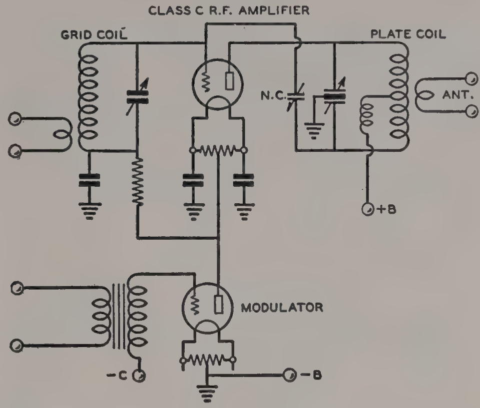


Figure 18.
SERIES PLATE MODULATION.

it is possible to secure quite good modulation up to about 90 per cent by applying modulation only to the plate, provided that the screen voltage and excitation are first run up as high as the tube will stand safely. Under these conditions, the screen tends to modulate itself to an extent, the screen voltage varying over the audio cycle as a result of the screen impedance increasing with plate voltage, and decreasing with a decrease in plate voltage.

The modulation transformer for plate-and-screen-modulation, when utilizing a dropping resistor, is similar to the type of transformer used for any plate-modulated phone. In Figure 17, the combined screen and plate current is divided into the plate voltage in order to obtain the class C amplifier load impedance. The audio power required to obtain 100 per cent sine-wave modulation is one-half the d.c. power input to the screen, screen resistor, and plate of the modulated r.f. stage.

Series Modulation. Another form of plate modulation is known as *series modulation,* in which the r.f. tube and modulator are in series across the d.c. plate supply, as shown in Figure 18.

Series modulation eliminates the modulation choke required in the usual form of Heising modulation. Although this system is capable of very good voice quality, the antenna coupling must be carefully adjusted simultaneously with the C bias in the modulator in order to maintain at least 20 per cent more plate voltage across the modulator than that which is measured from positive B to r.f. tube filament. It is difficult to obtain a high degree of modulation unless a portion of the total plate current is shunted by the r.f. tube through a resistor in series with a high-inductance choke coil. Series modulation is seldom used today except for television work.

The Doherty Linear Amplifier and the Terman-Woodyard Grid-Bias Modulated Amplifier

These two new-design amplifiers will be described collectively since they operate upon a very similar principle. Figure 19 shows a greatly simplified schematic diagram of the operation of both types. Both systems operate by virtue of a carrier tube (V_1 in both Figures 19 and 20) which supplies the unmodulated carrier, and whose output is reduced to supply negative peaks, and a peak tube (V_2 in Figures 19 and 20) whose function is to supply approximately half the positive peak of the modulation cycle and whose additional function is to lower the load impedance on the carrier tube so that it will be able to supply the other half of the positive peak of the modulation cycle.

The peak tube is enabled to increase the output of the carrier tube by virtue of an impedance inverting line between the plate circuits of the two tubes. This line is designed to have a characteristic impedance of one-half the value of load into which the carrier tube operates under the carrier conditions. Then a load of one-half the characteristic impedance of the quarter-wave line is coupled into the output. By experience with quarter-wave lines in antenna-matching circuits we know that such a line will vary the impedance at one end of the line in such a manner that the geometric mean between the two terminal impedances will be equal to the characteristic impedance of the line. Thus, if we have a value of load of *one-*

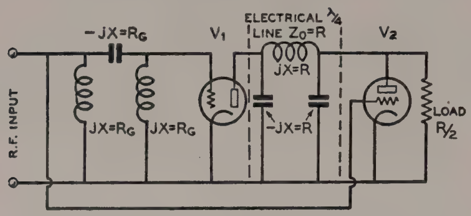


Figure 19.

half the characteristic impedance of the line at one end, the other end of the line will present a value of *twice* the characteristic impedance of the line to the carrier tube V_1.

This is the situation that exists under the carrier conditions when the peak tube merely floats across the load end of the line and contributes no power. Then, as a positive peak of modulation comes along, the peak tube starts to contribute power to the load until at the peak of the modulation cycle it is contributing enough power so that the impedance at the load end of the line is equal to R, instead of the R/2 that is presented under the carrier conditions. This is true because at a positive modulation peak (since it is delivering full power) the peak tube subtracts a negative resistance of R/2 from the load end of the line.

Now, since under the peak condition of modulation the load end of the line is terminated in R ohms instead of R/2, the impedance at the *carrier-tube* will be *reduced* from 2R ohms to R ohms. This again is due to the impedance inverting action of the line. Since the load resistance on the carrier tube has been reduced to half the carrier value, its output at the peak of the modulation cycle will be doubled. Thus we have the necessary condition for a 100 per cent positive modulation peak; the amplifier will deliver four times as much power as it does under the carrier conditions.

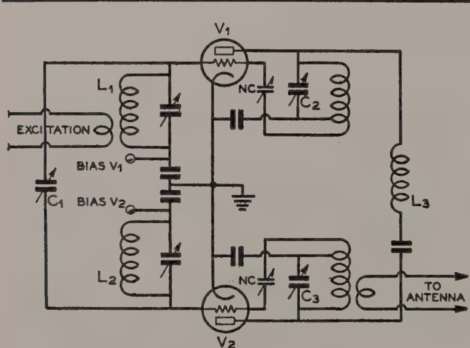


Figure 20.

On negative modulation peaks the peak tube does not contribute; the output of the carrier tube is reduced until on a 100 per cent negative peak its output is zero.

The Electrical Quarter-Wave Line. While an electrical quarter-wave line (consisting of a pi network with the inductance and capacity legs having a reactance equal to the characteristic impedance of the line) does have the desired impedance-inverting effect, it also has the undesirable effect of introducing a 90° phase shift across such a line. If the shunt elements are capacitances, the phase shift across the line leads by 90°; if they are inductances, the phase shift lags by 90°. Since there is an undesirable phase shift of 90° between the plate circuits of the carrier and peak tubes, an equal and opposite phase shift must be introduced in the exciting voltage to the grid circuits of the two tubes so that the resultant output in the plate circuit will be in phase. This additional phase shift has been indicated in Figure 19 and a method of obtaining it has been shown in Figure 20.

Comparison Between Linear and Grid Modulator. The difference between the Doherty linear and the Terman-Woodyard grid modulator is the same as the difference between a linear and a grid-modulated stage. Modulated r.f. is applied to the grid circuit of the Doherty linear with the carrier tube biased to cutoff and the peak tube biased to the point where it draws substantially zero plate current at the carrier condition. In the Terman-Woodyard grid modulated amplifier the carrier tube runs class C with comparatively high bias and high plate efficiency, while the peak tube again is biased so that it draws almost no plate current. Unmodulated r.f. is applied to the grid circuits of the two tubes and the modulating voltage is inserted in series with the fixed bias voltages. From one-half to two-thirds as much *audio* voltage is required at the grid of the peak tube as is required at the grid of the carrier tube.

High Operating Efficiencies. The resting carrier efficiency of the grid-modulated amplifier may run as high as is obtainable in any class C stage, 80 per cent or better. The resting carrier efficiency of the linear will be about as good as is obtainable in any class B amplifier, 60 to 65 per cent. The overall efficiency of the bias-modulated amplifier at 100 per cent modulation will run about 75 per cent; of the linear, about 60 per cent.

The effect of the quarter-wave line in the plate and grid circuits of the amplifier shown in Figure 20 is obtained by detuning the circuits enough to give the shunt element of the networks. At resonance, the coils L_1 and

L_2 in the grid circuits of the two tubes have each an inductive reactance equal to the capacitive reactance of the condenser C_1. Thus we have the effect of a pi network consisting of shunt inductances and series capacitance. In the plate circuit we want a phase shift of the same magnitude but in the opposite direction; so our series element is the inductance L_3 whose reactance is equal to the characteristic impedance desired of the network. Then the plate tank condensers of the two tubes C_2 and C_3 are increased an amount past resonance, so that they have a capacitive reactance equal to the inductive reactance of the coil L_3. It is quite important that there be no coupling between the inductors.

Although both these types of amplifiers are highly efficient and require no high-level audio equipment, they are difficult to adjust—particularly so on the higher frequencies—and it would be an extremely difficult problem to design a multi-band rig employing the circuit. However, the grid-bias modulation system has advantages for the high-power transmitter that may make some amateurs interested more than academically in the circuit. For those who are, discussion of the design and adjustment of the circuit has been given in recent issues of the technical radio magazines and in the *Proceedings* of the I.R.E.

Speech Equipment—Microphones

The microphone, which changes sound into electrical energy, usually consists of a diaphragm which moves in accordance with the compressions and rarefactions of the air called *sound waves*. The diaphragm then actuates some form of device which changes its electrical properties in accordance with the amount of physical movement.

If the diaphragm is very tightly stretched, the natural period of its vibration can be placed at a frequency which will be out of range of the human voice. This obviously reduces the sensitivity of the microphone, yet it greatly improves the uniformity of response to the wide range encountered for voice or musical tones. If the natural mechanically resonant period of the diaphragm falls within the voice range, the sensitivity is greatly increased near the resonant frequency. This results in distorted output, a familiar example being found in the old-type land-line telephone microphone.

A good microphone must respond equally to all voice frequencies; it must not introduce noise, such as hiss; it must have sufficient sensitivity to eliminate the need of excessive audio amplification; its characteristics should not vary with changes in temperature or humidity,

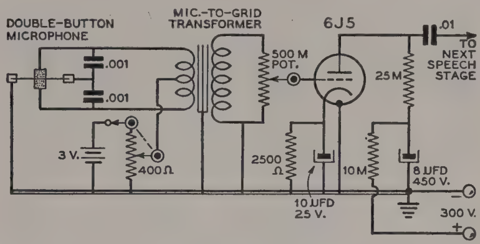


Figure 21.
SPEECH AMPLIFIER CIRCUIT FOR USE WITH A DOUBLE-BUTTON MICROPHONE.

and its characteristics should remain constant over a useful period of life.

The Carbon Microphone. Carbon microphones can be divided into two classes: (1) *single-button,* (2) *double-button.* The single-button microphone consists of a diaphragm which exerts a mechanical pressure on a group of carbon granules. These granules are placed behind the diaphragm between two electrodes, one of which is secured directly to the diaphragm and moves in accordance with the vibration of the diaphragm. This vibration changes the pressure on the carbon granules, resulting in a change of electrical resistance to current flowing between the electrodes, the direct current being supplied from an external source. The variation in resistance causes a change in the current which flows through the primary winding of a coupling transformer, thereby inducing a voltage in the secondary winding of this transformer; this voltage is then amplified by means of vacuum tubes.

Single-button microphones are useful for operation in portable transmitters because their sensitivity is greater than that of other types of microphones, thereby requiring less audio amplification to supply audio modulating power for the transmitter. The objectionable feature of the single-button microphone is its high hiss level. Another is that the diaphragm generally resonates within the voice range, resulting in mediocre tone quality. The better microphones of this type, however, are highly intelligible even though lacking somewhat in fidelity.

Double-Button Microphones. The double-button microphone has two groups of carbon granules arranged in small containers on either side of the diaphragm. This push-pull effect reduces the even-harmonic distortion, resulting in more intelligible modulation. The diaphragm is normally stretched to such an extent that its natural period may be as high as 8,000 cycles per second, which is beyond the

range of the human voice. This reduces the sensitivity of the microphone and greater audio amplification is needed to secure the same output as from a single-button carbon microphone. On the other hand, the tone quality from the double-button microphone is better, though the hiss is still present.

The cost of a double-button microphone is a satisfactory index of its performance when purchased from a reliable concern. The output from a *high-quality* two-button microphone is about 45 db below that of a standard single-button microphone.

Condenser Microphones. A condenser microphone has a better frequency response than a carbon microphone, and it does not produce a hiss. This type of microphone consists of a highly damped or stretched diaphragm mounted very close to a metal plate, but insulated from the plate. The movement of the diaphragm changes the spacing between the two electrodes, resulting in a change in electrical capacity. When a d.c. polarizing voltage is applied across the plates, an a.c. voltage will be generated when the diaphragm is actuated by reason of the change in capacity between the plates; this voltage can then be amplified by means of vacuum tubes. The diaphragm of a typical condenser microphone is made of duralumin sheet, approximately 1/1000 inch thick, with approximately the same spacing between the diaphragm and the rear heavy plate electrode. The output is approximately 75 db below an ordinary single-button carbon microphone with unstretched diaphragm.

The condenser microphone has a low output level, which necessitates at least two stages of preamplification, the first stage being located very close to the microphone. The output impedance is extremely high and the unit must, therefore, be well shielded in order to prevent r.f. and 60-cycle a.c. hum pickup. It is sensitive to changes in barometric pressure and humidity. More modern types of microphones are replacing the condenser type, although the latter are still often used.

Crystal Microphones. The crystal microphone operates on the principle that a change in dimensions of a *piezoelectric* material, such as a *Rochelle salt crystal,* generates a small a.c. voltage which can be amplified by means of vacuum tubes. No d.c. polarizing voltage or current or coupling transformer is required for the crystal type of microphone; thus, it becomes a very simple device to connect into an audio amplifier.

Crystal microphones can be divided into two classifications: (1) the diaphragm type, (2) the grille type.

The diaphragm type is relatively inexpensive, and consists of a semifloating diaphragm which subjects the crystal to deformation in accordance with the applied sound pressure. The fidelity is equal to that of most two-button carbon microphones, and there is no background noise or hiss generated in the microphone itself.

The grille type consists of a group of crystals connected in series or series-parallel, for the purpose of obtaining high electrical output

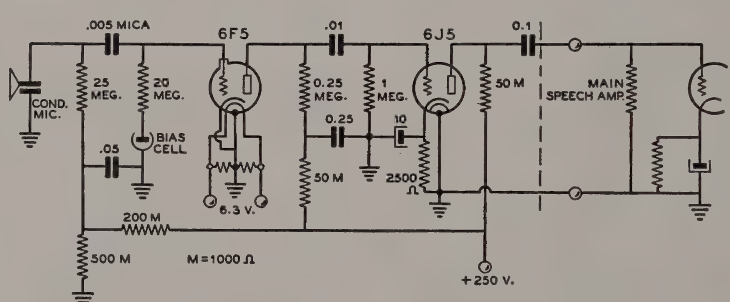


Figure 22.
CONDENSER MICROPHONE PRE-AMPLIFIER.
This pre-amplifier can be used to raise the level of the output of a condenser microphone to a point where it can be fed into the input of a normal speech amplifier. No transformer is needed to couple the 6J5 to the speech amplifier, unless it is desired to use a low-impedance line between them. The pre-amplifier should either be operated from a separate power supply or adequate decoupling should be used between the main speech amplifier and the pre-amplifier.

without aid of a diaphragm.

The output level varies between —55 db and —80 db for various types of crystal microphones. The grille type is less directional to sound pickup than most other types, and is capable of almost perfect fidelity. However, they have the disadvantage of a high thermal-agitation noise level.

Velocity or Ribbon Microphones. The inductive or ribbon-type microphone has a thin, corrugated, metal strip diaphragm which is loosely supported between the poles of a horseshoe magnet. A minute current is induced in this strip when it moves in a magnetic field, and this current can be fed to the primary of a step-up-ratio transformer of high ratio because of the very low impedance of the ribbon.

The microphone output must be amplified by means of a very high gain preamplifier, because the output level of the older types of ribbon microphones is —100 db and even the newer ones are around —85 db. The inductive type of microphone is rugged and simple in construction. Unfortunately, it cannot be used for close talking without overemphasizing the lower frequencies. It is a velocity, rather than a pressure-operated, microphone, and should therefore be placed at least 2 feet from the source of sound. It is very sensitive to a.c. hum pickup, and this is one of the principal reasons why it is not widely used in amateur practice.

The impedance of the ribbon is so low that it is difficult to design a ribbon-to-grid transformer with good fidelity. Therefore, for best quality, two transformers are usually used in cascade: ribbon-to-200 ohms and 200 ohms-to-grid.

The Dynamic Microphone. The dynamic (moving coil) type of microphone operates on the same principle as the inductive microphone. A small coil of wire, actuated by a diaphragm, is suspended in a magnetic field, and the movement of the coil in this field generates an alternating current. The output impedance is approximately 30 ohms as against approximately 1 ohm for the ribbon type of microphone. The output level of the high fidelity types is about —85 db, the level varying with different makes. The output level of the p.a. types is somewhat higher, and the fidelity is almost as good. This type of microphone is quite rugged, but has the disadvantage of picking up hum when used close to any power transformers.

An inexpensive and very satisfactory dynamic microphone for amateur transmitters can be made from a small, permanent-magnet type, dynamic loudspeaker. One of the newer 5-inch types with alloy magnet will give surprising fidelity at relatively high output level.

A shielded cable and plug are essential to prevent hum pickup. The unit can be mounted in any suitable type of container. The circuit diagram is shown in Figure 25.

Directional Effects. Crystal microphones, as well as those of some other types, can be mounted in a spherical housing with the dia-

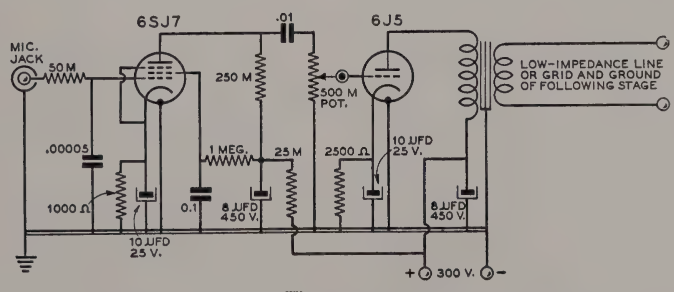


Figure 23.

RECOMMENDED SPEECH AMPLIFIER INPUT CIRCUIT.

This is a simple and conventional speech amplifier circuit for operating out of a crystal, high-impedance dynamic, or other low-level microphone. The voltage gain will be in the vicinity of 2300, which means that the amplifier will have a gain of about 67 db. With a crystal mike with output of −50 db the output of the 6J5 will be about plus 17 db; this is ample to drive a pair of 2A3's as a driver for a class B modulator. In this case a push-pull input transformer would be used in the plate circuit of the 6J5. If it is desired to feed a low-impedance line, a plate-to-line transformer should be used in the 6J5 plate circuit.

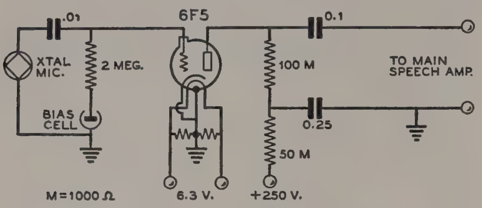


Figure 24.
PRE-AMPLIFIER—GAIN APPROXIMATELY 35 DB.
This pre-amplifier can be used either with a crystal microphone as shown, or with a dynamic microphone by changing the resistance network in the input circuit to a transformer of the correct design to match the dynamic microphone to the grid.

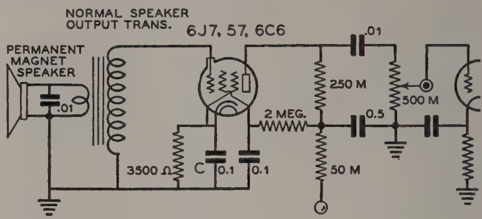


Figure 25.
DYNAMIC MICROPHONE INPUT AMPLIFIER.
A low-cost microphone arrangement using a midget permanent-magnet dynamic speaker as the microphone and its output transformer as the coupling transformer between it and the grid of the first audio stage.

phragm oriented horizontally in order to secure a non-directional effect. Decidedly directional effects may be required, on the other hand, and microphones for this purpose are commercially available.

Speech Amplifiers

That portion of the audio channel between the microphone or its preamplifier and the power amplifier or driver stage can be defined as the *speech amplifier.* It consists of from one to three stages of *voltage amplification* with resistance, impedance, or transformer coupling between stages. The input level is generally about −50 db in the case of a speech amplifier designed for a double-button carbon microphone or preamplifier input. The input level is approximately −70 db when the speech amplifier is designed for operation from a diaphragm-type crystal microphone. Some conventional speech amplifier circuits are shown in the preceding pages. Other speech amplifier circuits are shown in the chapter on *Speech and Modulation Equipment.*

It is possible to dispense with the preamplifier with certain types of low-level microphones by designing the speech amplifier input to work at −100 db or so, but it is better practice and entails less constructional care if a speech amplifier with less gain is used, in conjunction with a preamplifier to make up the required overall amplification. Less trouble with hum and feedback will be encountered with the latter method.

Designing a speech amplifier to work at −70 db is comparatively easy, as there is little trouble from power supply hum getting into the input of the amplifier by stray capacitive or inductive coupling.

Amplifier Gain. The power gain in amplifiers, or the power loss in attenuators, can be conveniently expressed in terms of db *units,* which are an expression of ratio between two power levels.

A formula for the calculation of db gain or loss is here given:

$$\text{db} = 10 \times \log_{10} \frac{P_2}{P_1}$$

Since power is equal to the product of voltage times current when the power factor is unity, db units can be used to express voltage gain. In this case the formula is:

$$\text{db} = 20 \times \log_{10} \frac{E_1}{E_2}$$

This provides a useful means for computing the overall voltage gain of a preamplifier and the speech amplifier. When adding the gain of several stages, the db units are added or subtracted, which greatly simplifies the calculations.

For example: if a preamplifier has 35 db gain, and the speech amplifier has 65 db gain, the total gain is 35 + 65, which equals 100 db. One hundred db corresponds to a voltage gain of 100,000 times. Thus, for example, if the microphone level is −100 db the speech amplifier output will be −100 db +100 db, or zero db level. Zero level corresponds to a power level of 6 milliwatts.

In order to obtain 60 watts of audio power output, a power gain of 6,000 times will be required, which corresponds to a power gain of approximately 38 db. This amplification can be considered as part of the main power amplifier or modulator, or as part of the speech amplifier, depending upon the particular trans-

mitter under consideration. The important point to remember is that power ratios use the expression: $10 \times \log$, whereas voltage gain between similar impedances is computed by the expression: $20 \times \log$.

Let us take a typical example of a radiophone transmitter with a class C amplifier input of 200 watts. For 100 per cent plate modulation, the audio power requirement is 100 watts. This corresponds to a db power level of + 42 db. Zero db level is 6 milliwatts or .006 watt. (Refer to db power table in Chapter 28.)

Therefore, the formula:

$$db = 10 \times \log_{10} \frac{P_1}{P_2}$$

$$10 \times \log_{10} \frac{100}{.006} = 42$$

The amateur may desire to use a cell (grille) type crystal microphone which is rated at −70 db for average sound levels. This extremely low output must be brought up to a value of 100 watts or +42 db. The total gain required will be 112 db.

No preamplifier would be necessary, because this amount of gain can be built into a good speech amplifier and modulator. A typical audio channel which meets these requirements is shown in the skeleton circuit, Figure 26.

The first speech amplifier consists of a 6SJ7 connected as a high-gain pentode, resistance-coupled to a 6J5 speech amplifier which, in turn, is coupled through a step-up transformer into a 6F6 tube which operates as a triode. The latter is connected to a push-pull 45 class AB driver for the final power amplifier or modulator, consisting of a pair of 35T's.

The 6SJ7 stage is capable of producing a voltage amplification of 100 times, which corresponds to 40 db.

$$db = 20 \times \log_{10} \frac{100}{1} = 40$$

The 6J5 and 6F6 triodes with a 3-to-1 step-up interstage transformer will produce a voltage gain of 240.

$$db = 20 \times \log_{10} \frac{240}{1} = 47$$

Actually, the db voltage gain must be measured between like impedances in order to be correct.

The total speech amplifier gain is 40 + 47, which equals 87 db. If the output level of the microphone is −70 db, the output level of the 42 triode will be 87 − 70, which equals +17 db. This level corresponds to approximately 300 milliwatts, which is well within the rating of a 6F6 triode driver, and is sufficient to drive the 45 tubes in class AB.

$$17 = 10 \times \log_{10} \frac{P}{.006}$$

Therefore, P equals 0.3 watt or 300 milliwatts.

$$db = 10 \times \log_{10} \frac{100}{0.3} = 25$$

This can be checked by subtracting 17 from 42, which is 25 db, the power gain between the grids of the 45 tubes and the output of the class B modulator.

With 0.3 watt input to the 45 stage, 9 watts of output can be obtained.

$$db = 10 \times \log_{10} \frac{9}{0.3} = 15$$

The power gain through the 45 stage is 15 db, leaving a power gain of 10 in the 35T class B stage. More power gain could be secured in the 35T stage, thus requiring less gain in the 45 driver stage, and therefore the class B input transformer could have a greater step-down ratio than in the case of a circuit design

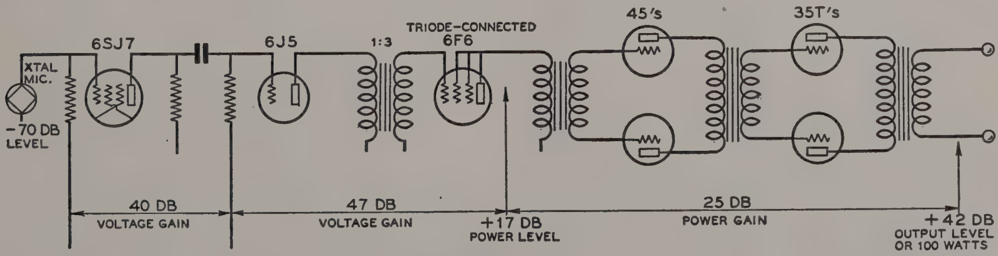


Figure 26.
CALCULATION OF THE GAIN OF THE AUDIO CHANNEL OF A RADIOPHONE TRANSMITTER.

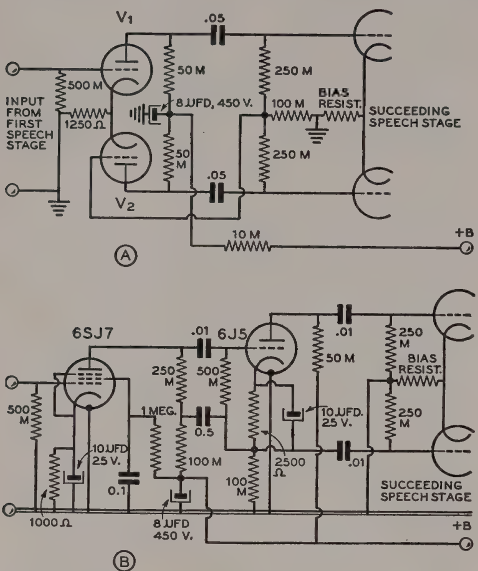


Figure 27.
RECOMMENDED PHASE-INVERTER CIRCUITS.

(A) shows the "floating paraphrase" circuit which is the best for all ordinary applications. (B) shows an excellent complete front end for a speech amplifier which will deliver ample voltage output to drive any of the conventional power audio tubes when a conventional crystal or high-impedance dynamic microphone is used on the input. The circuit shown at (A) is quite flexible and is capable of considerable change to suit different circuit conditions. That shown at (B) should be used as is without change if its excellent operating characteristics are to be retained. Both circuits are fully described in the accompanying text.

in which no leeway in voltage and power gain is provided for.

Phase Inverters. Quite frequently in the design of a speech amplifier it is desirable to go from a single-ended microphone input stage, or from the single-ended stage following the input amplifier, into a push-pull power output stage, or a push-pull driver for class B grids. Until recent years a push-pull input transformer was employed to obtain the 180°-out-of-phase voltages necessary for the grids of the output tubes. But good quality push-pull input transformers are expensive and have a tendency toward picking up inductive hum and introducing phase shift. The present trend is toward the use of a *phase-inverter* circuit between the single-ended low-level stage and the push-pull output amplifier.

Figure 27 shows two phase-inverter circuits which have proved to give excellent results in all normal types of applications. Both make use of degenerative feedback to stabilize and equalize the voltages developed across the two halves of the output circuit. The circuit shown at (A) can be used with any two tubes of the low-power types usually employed in low-level audio work. One of the most satisfactory arrangements is to use a dual tube for both V_1 and V_2. The 6N7, 6F8-G, 6SC7, 6SN7, 7F7, and 6Z7-G twin triodes all are suitable for this application. The voltage gain of this phase inverter from the grid of V_1 to the two grids of the succeeding speech stage is slightly less than *twice* the actual gain of V_1. V_1 and V_2 need not be biased from the same cathode resistor; they may each have a separate cathode resistor (by-passed or not by-passed, as desired) and degenerative feedback from the output of the amplifier may be fed back into the cathode of V_1 (but *not* into V_2). The voltage which appears on the grid of V_2 arises from the unbalance in the output voltages delivered by the two phase-inverter tubes. Hence, the higher the gain of the tube at V_2 the less will be the difference between the voltages fed to the two output stage grids. In any case, if the gain of V_2 is above 15, the voltage appearing on the grid fed by V_2 will be at least 94 per cent of that appearing at the other grid.

The circuit shown at (B) of Figure 27 has a total voltage gain from the input grid of the 6SJ7 to the two grids of the push-pull speech stage of about 2300. This is ample gain to operate from a source such as a crystal microphone or pickup into the grids of a pair of 6A3's, 6V6's, or 6L6's. The 6SJ7 stage gives a gain of about 150, while the 6J5 gives a total gain of about 14, or about 7 for each output tube grid. This circuit is unique among cathode-follower phase-inverter circuits in that the full gain of the cathode follower tube (6J5) is obtained, although this gain is split, of course, between the two grid circuits of the following stage. Slight adjustments in the value of the 100,000-ohm resistor in the cathode circuit of the 6J5 will allow exactly equal and opposite voltages to be obtained on the grids of the succeeding stage.

Class A Triode Amplifier Calculations. In the design of a speech amplifier or modulator it is often desirable or necessary to be able to determine beforehand exactly how a particular class A triode amplifier stage is going to perform under certain specified operating conditions. Figure 28 gives four very useful formulas for making certain calculations of the operation of a triode amplifier stage. The particular stage may be a 6J5 triode operating with 0.1 volt output, or it may be a 10-kilo-

Figure 28.

FORMULAS FOR CALCULATING THE OPERATING CONDITIONS OF A TRIODE CLASS A AMPLIFIER STAGE.

ZERO SIGNAL BIAS

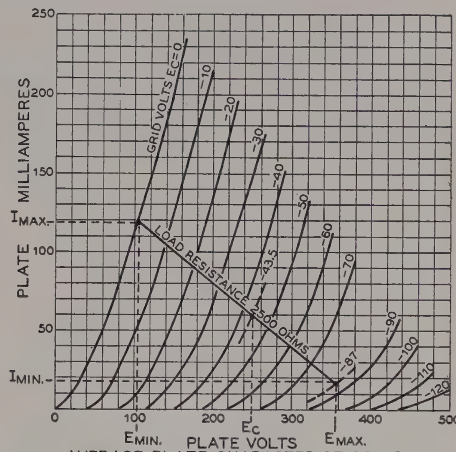
$$E_C = -\frac{0.68\,E_B}{\mu}\ \text{VOLTS}$$

LOAD RESISTANCE

$$R_L = \frac{E_{MAX.} - E_{MIN.}}{I_{MAX.} - I_{MIN.}}\ \text{OHMS}$$

POWER OUTPUT

$$P_O = \frac{(I_{MAX.} - I_{MIN.})\,(E_{MAX.} - E_{MIN.})}{8}\ \text{WATTS}$$



AVERAGE PLATE CHARACTERISTICS—2A3

$\mu = 4.2$ $R_P = 800$ OHMS
PLATE DISSIPATION = 15 WATTS

SECOND HARMONIC DISTORTION

$$D_2 = \frac{\dfrac{(I_{MAX.} + I_{MIN.})}{2} - I_O}{I_{MAX.} - I_{MIN.}} \times 100\ \text{PERCENT}$$

watt Heising modulator—the formulas apply equally well. But it is necessary, in any case, that the tube in question (or tubes if they are operating in push-pull) be operating class A with substantially constant plate current, and that it be a triode.

The plate characteristics of a 2A3 have been given in Figure 28 along with a load line already drawn so as to offer an example of the method of calculation. The first step in making a calculation is to decide upon the plate voltage which will actually appear at the plate of the tube. For this example, suppose we take 250 volts. Then the approximate static grid bias may be obtained from the formula marked "Zero Signal Bias."

$$E_c = 0.68 \times 250/4.2 = 40.5 \text{ volts}$$

Since the 2A3 is a low-μ low-R_p triode it is advisable to raise the static grid bias a few per cent to, say, 43.5 volts. Since the plate dissipation rating of the 2A3 is 15 watts, if we divide this value by the plate voltage we can obtain the value of the maximum plate current.

$$I_p = 15/250 = 0.060 \text{ ampere or } 60 \text{ ma.}$$

Checking these three values on the plate characteristics we find that they correspond exactly. This point on the curve is called the operating point. If the operating point chosen by the first approximation of grid bias exceeds any of the tube ratings, the operating point should be moved slightly until all values are within the tube ratings. After the operating point has been determined (—43 volts grid bias, 250 volts plate, 60 ma. plate current, in this case) the load line can be drawn to extend from zero grid bias, through the operating point to twice the operating bias (—87 volts, in this case). For 5 per cent distortion the load line should extend in a maximum ratio of 11 units upward to 9 units downward. If the ratio of the positive swing to the negative swing is more than 11 to 9, the distortion will be greater than 5 per cent; if it is less than 11 to 9 (closer to a 1-to-1 ratio) the full output capabilities of the tube will not be obtained, although the distortion will be less than 5 per cent.

When the load line has been drawn, with the above taken into consideration, the load resistance (which is determined graphically by the slope of the line) can be determined by the second formula given. In the case shown it is:

$$R_L = (357 - 107)/(0.118 - 0.018) = 250/.100 = 2500 \text{ ohms}$$

The value determined is the value of load resistance which should be reflected by the secondary of the transformer to the plate of the tube. Then the power output of the tube operating under these conditions may be determined by the third formula given. In this case it is:

$$P_o = (0.118 - 0.018)\,(357 - 107)/8 = 25/8 = 3.12 \text{ watts}$$

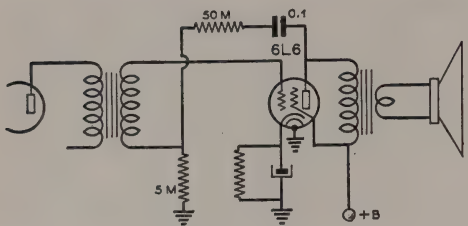


Figure 29.
INVERSE FEEDBACK FOR A SINGLE-STAGE BEAM-TUBE AMPLIFIER.

If the ratio of the two sides of the load line (either side of the operating point) is 11 to 9, the distortion will be 5 per cent. However, if the ratio is any value other than 11 to 9, the distortion will be more or less than 5 per cent. In any case, the distortion may be calculated by use of the last formula given. In this case it will be:

$$D_2 = \frac{(0.118 + 0.018)/2 - .060}{0.118 - 0.018} = 8/100 = 8 \text{ per cent}$$

If the value of load resistance, power output, and distortion obtained are not satisfactory, it is merely necessary to draw new load lines until the desired values are obtained for the operating conditions.

Modulators. A *modulator* supplies audio power to the particular r.f. stage in the transmitter which is being modulated. A speech amplifier does not deliver sufficient power output for modulating a conventional form of r.f. stage delivering more than a very few watts power. The modulator is an audio amplifier which delivers ample power output for completely modulating the d.c. input to the modulated stage. Power requirements of audio amplifiers vary from a fraction of a watt up to 500 watts, for amateur purposes. Low-power transmitters of the grid-modulated or suppressor-grid-modulated types require less than 1 watt of audio power, whereas a 1-kw. plate-modulated phone transmitter requires 500 watts of audio power for 100 per cent sine-wave modulation.

Figure 34 is a simple diode rectifier which incorporates a phase-reversing switch, which must be thrown to that position, which will cause a slight reduction in speech amplifier gain. The actual gain of the spech amplifier can be increased by means of the manual gain control. The undesired noise or hum which is audible in the phone monitor will generally be reduced with the correct adjustment of the r.f. pickup coil and phase-reversing switch. Once adjusted, no additional changes are necessary, unless the transmitter power output or frequency is varied.

In Figure 35, a type 84 rectifier tube is connected so that one side serves as an inverse feed-back rectifier, and the other side is a standard overmodulation indicator and 'phone monitor.

The circuits in Figures 36 and 37 show methods of connection from the feedback rectifier in to the speech amplifier.

The Diode Feedback Rectifier. The diode feedback rectifier rectifies the carrier, and any hum or noise modulation on the carrier appears as an audio voltage across the 100,000-ohm feedback control to the grid of the speech amplifier. A portion of this voltage is fed back into the speech amplifier so as to be out of phase, and thus buck out the hum or noise in the output of the radio transmitter. This may actually introduce distortion in a portion of the speech amplifier in which there is otherwise none present (commonly spoken of as being within the feedback "loop") but the final result is that the distortion or hum is reduced in the carrier signal of the radiotelephone transmitter.

Limitations of Degenerative Feedback. It might seem from the above information that degenerative feedback is a "cure-all" for any type of distortion, hum, or non-linearity which may arise in a speech amplifier, modulator, or transmitter. This is most certainly not the case. In fact, the limitations of feedback are such that it must be very carefully applied if the greatest results are to be obtained.

The primary enemy of degenerative feedback is phase shift. Were it not for phase shift, degenerative feedback might be the cure-all that it appears to be on first inspection. But it must be considered that while the feedback voltage may be 180° out of phase with the incoming signal at some intermediate frequency, the phase shift in the audio amplifier or modulator at some audio frequency from 8000 to 25,000 cycles may be such that the feedback voltage will now be *in phase* with the incoming voltage. When this occurs the amplifier will, of course, oscillate if there is sufficient gain. Hence, the most important consideration in designing a circuit around which degenerative feedback is to be placed, is that phase shift in every portion of the circuit be kept to a minimum. Since all audio transformers introduce a certain amount of phase shift, it is seldom possible to include more than two audio transformers within a feedback loop —frequently one is all that can be used. Any type of step-up transformers should be

avoided; all transformers used should be 1-to-1 ratio or step down.

Classes of Modulators. Class A amplifiers are suitable for low-power grid-modulated, or suppressor-modulated phone transmitters; class AB audio amplifiers for high-power grid-modulated or for low-power plate-modulated phones, and class B audio amplifiers for most economical operation of transmitters in which the audio requirements are greater than about 50 watts. Class AB or class B modulators require a *driver* stage, which can be considered part of the modulating system proper rather than part of the speech amplifier. The complete modulator essentially consists of a device for converting speech-amplifier output *voltage* into audio *power.*

Complete information on receiver and transmitter type tubes for modulator service, as well as for any other portion of a radio-telephone transmitter, will be found in the material on receiving and transmitting tubes, Chapters 5 and 10.

Degenerative Feedback

A system of taking energy from the output of an amplifier or transmitter, and feeding it back into the input circuit 180° out of phase with the incoming voltage, has come into quite wide usage in recent years. Inverse feedback or degeneration, as it is called, allows greatly improved operation of audio amplifiers and radiophone transmitters to be obtained. It has been found that the proper application of degeneration in an amplifier can be made to reduce greatly the harmonic distortion and otherwise to improve the fidelity. The inclusion of inverse feedback causes a reduction in the gain of an amplifier in direct proportion to the reduction in noise and distortion. This can be offset by the addition of a stage of speech amplification. The disadvantage of the additional stage of amplification is far more than compensated for by the reduction in three kinds of distortion commonly known as frequency distortion (change in gain with respect to frequency), harmonic distortion, and delay or phase distortion, and by the reduction of any noise and hum arising within the feedback loop.

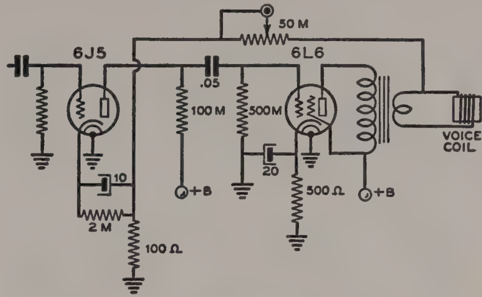


Figure 30.
INVERSE FEEDBACK AROUND A TWO-STAGE AMPLIFIER.

The Inverse Feedback Principle. The principle involved in inverse feedback systems is to select a portion of the amplifier output voltage and feed it back into one of the previous circuits, exactly out of phase with the input voltage. In Figure 29, a simple method of applying inverse feedback to an audio amplifier is shown. With the values of resistance as indicated, the reverse feedback is approximately 10 per cent. This reduces the gain of the audio amplifier; however, it still has approximately twice the sensitivity of a triode amplifier with similar plate circuit character-

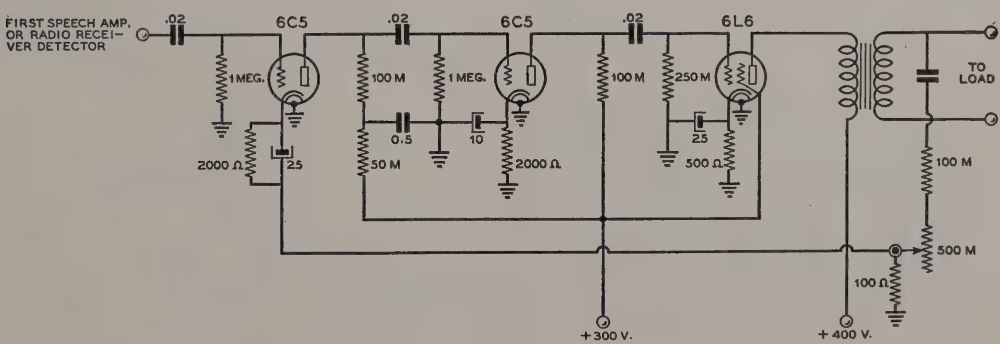


Figure 31.
THREE-STAGE DEGENERATIVE FEEDBACK AMPLIFIER.

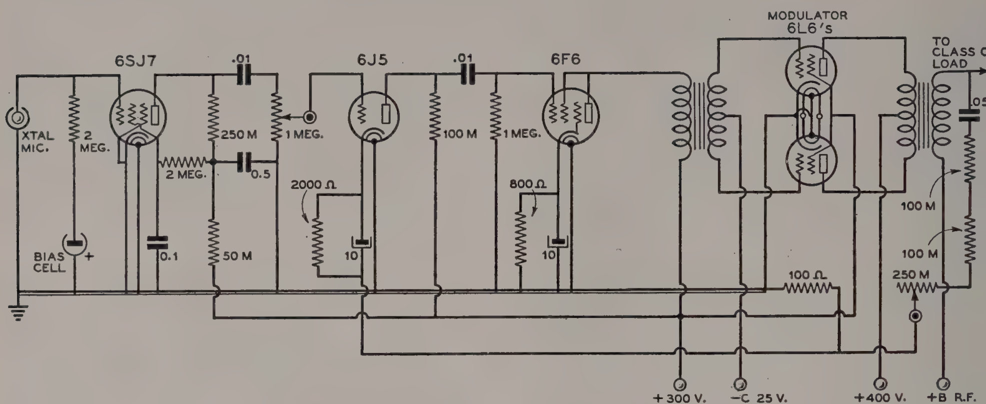


Figure 32.

45-WATT BEAM-TUBE MODULATOR WITH DEGENERATIVE FEEDBACK.

istics. The plate circuit impedance of the 6L6 is greatly reduced, an advantage when working into a loudspeaker (because a loudspeaker is not a constant impedance device).

Inverse feedback can be applied in a somewhat different manner, as shown in Figure 30, for a two-stage amplifier. This method is particularly desirable, in that feedback produces better results when the feedback circuit is connected from the output back to the grid of one of the *preceding* amplifier stages.

The polarity of the secondary winding of the output transformer, in all cases where the feedback connection is made to the secondary, should be that which will produce degeneration and reduction in amplifier gain, rather than regeneration and howl or increase of gain.

The circuits in Figures 31 and 32 indicate methods for applying inverse feedback to three stages of amplification. These two systems are suitable for operation as speech amplifiers and modulators for grid-modulated radio-telephones, or low-power plate-modulated transmitters. The 100-ohm cathode resistor should be located as near as possible to the 6C5 tube cathode terminal in order to prevent undesirable pickup and feedback at frequencies other than those desired.

Parallel Inverse Feedback Circuit. Figure 33 shows a particularly simple and effective means of obtaining degenerative feedback around a pentode or beam tetrode output stage. The distortion at all output levels of the 6L6 amplifier stage is greatly reduced, and the permissible power output, before serious distortion starts to occur, is increased from about 5 watts without feedback to about 6.5 watts with the feedback circuit shown. The circuit consists simply of a high-gain audio stage, using a tube with high plate impedance, coupled to a beam-tube or pentode output stage. Degenerative feedback is accomplished by the inclusion of the resistor in Figure 33

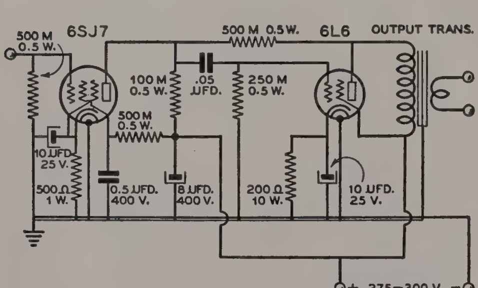


Figure 33.

6.5-WATT PARALLEL FEEDBACK AMPLIFIER-MODULATOR.

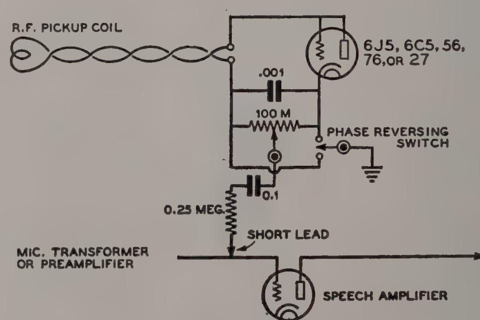


Figure 34.

DIODE CARRIER RECTIFIER WITH FEEDBACK TO AUDIO STAGE.

from the plate of the output tube to the plate of the 6SJ7.

Since the plate impedance of the 6L6 is lowered by the addition of feedback around it, the correct value of load impedance is 2500 ohms. The gain of the combined two-tube circuit is intermediate between the value required for excitation of the 6L6 alone and the value required with a 6SJ7 amplifier without feedback in front of it; about 1 volt peak is required at the grid of the 6SJ7 for full output from the 6L6. The circuit is satisfactory for use as a low-power grid or plate modulator, as a driver for a class B stage, or to operate a speaker. A speech amplifier using this circuit is given in Chapter 14.

R. F. Inverse Feedback. Modulation distortion, noises, and hum level which are present on the carrier of a radiotelephone station can be reduced by inverse feedback applied as in many broadcast transmitters, but modified for amateur applications. The method consists of rectifying a small amount of the carrier signal and feeding back the audio component in reverse phase into some part of the speech amplifier. This arrangement will reduce the hum level and improve the voice quality of most amateur radiotelephone transmitters.

The amount of inverse feedback that can be applied in this manner will depend upon the available amount of excess speech amplification, and the degree to which it can be carried without oscillation. The process of inverse feedback is to utilize voltages 180° out of phase over the band of frequencies of operation. Sometimes the feedback voltage may be considerably less than 180° out of phase for frequencies outside of the voice range, resulting in oscillation above the audible range, and the amount of feedback which can be applied is limited by this effect.

Two inverse feedback rectifier circuits are shown in Figures 34 and 35.

Other things which must be carefully avoided within the feedback loop are small values of interstage coupling condensers, any sort of shunting condensers such as a plate by-pass on a modulated stage, and large values of series resistance anywhere within the feedback loop. If there should arise any case of oscillation caused by too large a value of series resistor in the feedback circuit proper, this trouble can often be cured by shunting the series resistor with a very small value of mica condenser—0.00004 μfd. or so. However, in a case where it is impossible to eliminate oscillation in a circuit employing degenerative feedback, it is always possible to eliminate the difficulty by reducing the amount of feedback. In a circuit with a large amount of phase change with frequency, it may be necessary to reduce the feedback to an amount so small that it may as well be eliminated. This is the condition which usually arises when it is attempted to place degenerative feedback around a plate modulated transmitter using class B modulators. Degenerative feedback may be used satisfactorily from the rectified carrier back to the audio, in transmitters using Heising plate modulation, and suppressor or control-grid modulation. The system is especially suited to application in transmitters using grid-bias modulation.

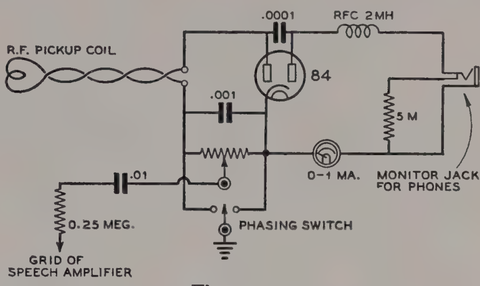


Figure 35.
DIODE CARRIER RECTIFIER FOR FEEDBACK WITH MONITORING CIRCUIT.

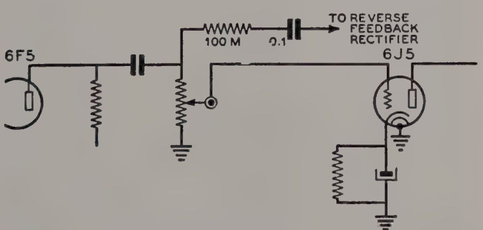


Figure 36.
PARALLEL COUPLING OF FEEDBACK INTO SPEECH AMPLIFIER.

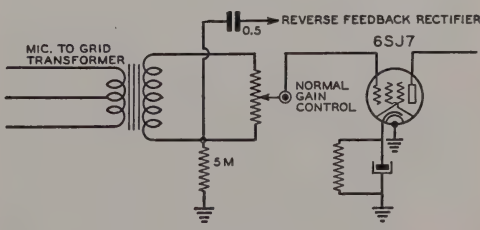


Figure 37.
SERIES COUPLING OF FEEDBACK ENERGY INTO SPEECH SYSTEM.

Automatic Modulation Control and Automatic Peak Limiting

It is possible to increase the average modulation level without danger of overmodulation, by designing the speech amplifier to have a non-linear amplification above a threshold value corresponding to approximately 80 per cent modulation. In other words, the gain of the amplifier is constant until a signal is impressed upon it that would ordinarily modulate the transmitter over 80 per cent; then the gain of the amplifier goes down rapidly as the input signal is increased.

To increase the modulation percentage in a conventional transmitter from 80 per cent to 100 per cent requires an increase in the input signal of 2 db. Broadcast stations commonly employ a compressor or peak limiter which requires 5 db increase in the audio input voltage to the amplifier in order to raise the modulation from 80 to 100 per cent. This gives 3 db compression, and permits running of the gain control, without danger of overmodulation, at a setting 3 db higher than would otherwise be possible. This is equivalent to doubling the transmitter power.

Somewhat more than 3 db compression can be employed in a voice transmitter designed for communication work, but an attempt to use too much compression will result in distortion so great as to affect the intelligibility.

Automatic modulation control is similar to a peak-limiting audio amplifier in effect, though the method of accomplishing the compression is somewhat different. In the a.m.c. system, the output of the modulator itself is used to actuate the compression circuit, and it is somewhat more positive in action and easier to adjust. The chief disadvantage of the latter system is that it can be used only with plate modulation, while a peak-limiting a.f. amplifier can be used with either plate- or any type of grid-modulation.

Practical application of peak-limiting and a.m.c. systems will be found in the chapter *Speech and Modulation Equipment.*

CHAPTER NINE

Frequency Modulation

To experimentally inclined amateurs, building and operating frequency modulation (f.m.) equipment offers much in the way of enjoyment and instruction. In this chapter the various points of difference between f.m. and amplitude modulation transmission and reception will be discussed and the advantages of f.m. for certain types of communication will be pointed out. Since the distinguishing features of the two types of transmission lie entirely in the modulating circuits at the transmitter and in the detector and limiter circuits in the receiver, these parts of the communication system will receive the major portion of attention.

Modulation. As previously described in this book, *modulation* is the process of altering a radio wave in accordance with the intelligence to be transmitted. The nature of the intelligence is of little importance as far as the process of modulation is concerned; it is the *method* by which this intelligence is made to give a distinguishing characteristic to the radio wave which will enable the receiver to convert it back into intelligence that determines the type of modulation being used.

Figure 1 is an oscillogram of an r.f. carrier amplitude modulated by a sine-wave audio voltage. After modulation the resultant modulated r.f. wave is seen still to vary about the zero axis at a constant rate, but the strength of the individual r.f. cycles is proportional to the amplitude of the modulation voltage.

In Figure 2, the carrier of Figure 1 is shown frequency modulated by the same modulating voltage. Here it may be seen that modulation voltage of one polarity causes the carrier frequency to decrease, as shown by the fact that the individual r.f. cycles of the carrier are spaced farther apart. A modulating voltage of the opposite polarity causes the frequency to increase, and this is shown by the r.f. cycles being squeezed together to allow more of them to be completed in a given time interval.

Figures 1 and 2 reveal two very important characteristics about amplitude- and frequency-modulated waves. First, it is seen that while the amplitude (power) of the signal is varied in a.m. transmission, no such variation takes place in f.m. In many cases this advantage of f.m. is probably of equal or greater importance than the widely publicized noise reduction capabilities of the system. When 100 per cent amplitude modulation is obtained, the average power output of the transmitter must be increased by 50 per cent. This additional output must be supplied either by the modulator itself, in the high-level system, or by operating one or more of the transmitter stages at such a low output level that they are capable of producing the additional output without distortion, in the low-level system. On the other hand, a frequency modulated transmitter requires an insignificant amount of power from the modulator and needs no provision for increased power output on modulation peaks. All of the stages between the oscillator and the antenna may be operated as high-efficiency class B or class C amplifiers or frequency multipliers.

The second characteristic of f.m. and a.m. waves revealed by Figures 1 and 2 is that both types of modulation result in distortion of the r.f. carrier. That is, after modulation, the r.f. cycles are no longer sine waves, as they would be if no frequencies other than the fundamental carrier frequency were present. It may be shown in the amplitude modulation case illustrated, that there are only two additional frequencies present, and these are the familiar "side frequencies," one located on each side of the carrier, and each spaced from the carrier by a frequency interval equal to the modulation frequency. In regard to frequency and amplitude, the situation is as shown in Figure 3. The strength of the carrier itself does not vary during modulation, but the strength of the side frequencies depends upon the percentage of modulation. At 100 per cent modulation the power in the side frequencies is equal to half that of the carrier.

Under frequency modulation, the carrier wave again becomes distorted, as shown in Figure 2. But, in this case, many more than

two additional frequencies are formed. The first two of these frequencies are spaced from the carrier by the modulation frequency, and the additional side frequencies are located out on each side of the carrier and are also spaced from each other by an amount equal to the modulation frequency. Theoretically, there are an infinite number of side frequencies formed, but, fortunately, the strength of those beyond the frequency "swing" of the transmitter under modulation is relatively low.

One set of side frequencies that might be formed by frequency modulation is shown in Figure 4. Unlike amplitude modulation, the strength of the component at the carrier frequency varies widely in f.m., and it may even disappear entirely under certain conditions. The variation of strength of the carrier component is useful in measuring the amount of frequency modulation, and will be discussed in detail later in this chapter.

One of the great advantages of f.m. over a.m. is the reduction in noise at the receiver which the system allows. If the receiver is made responsive only to changes in frequency, a considerable increase in signal-to-noise ratio is made possible through the use of f.m., when the signal is of greater strength than the noise. The noise reducing capabilities of f.m. arise from the inability of noise to cause appre-

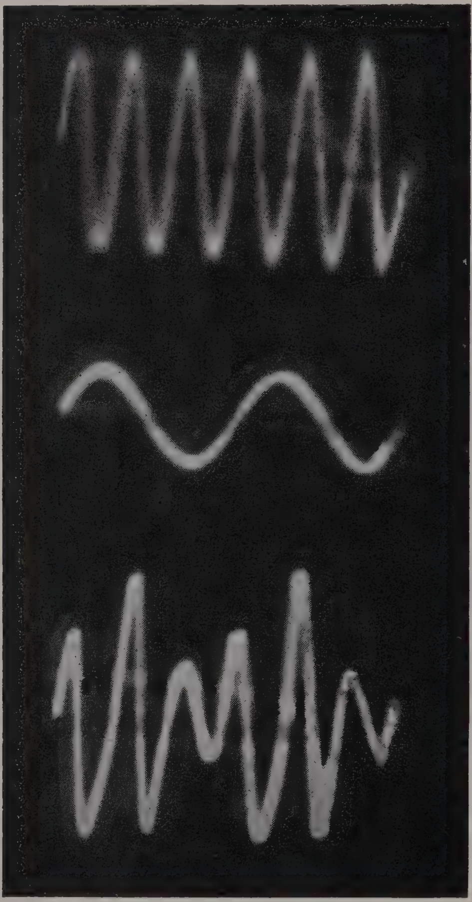


Figure 1.

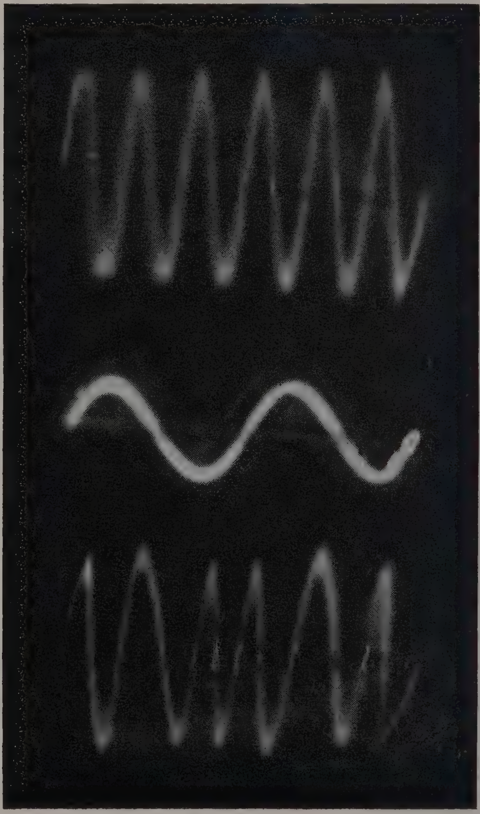
AMPLITUDE MODULATION.

Oscilloscope pattern of an amplitude-modulated wave. The unmodulated carrier is shown at the top, modulating wave at the center, and the resultant modulated wave at the bottom.

Figure 2.

FREQUENCY MODULATION.

The unmodulated carrier wave (top) and the modulating wave (center) are the same as those shown in Figure 1. Note the difference in the wave form of the resultant modulated wave (bottom).



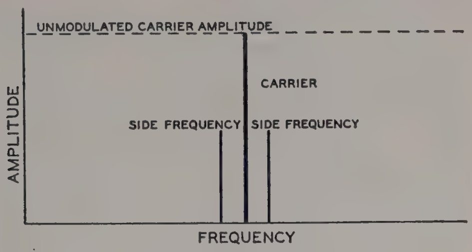


Figure 3.
A.M. SIDE FREQUENCIES.
For each a.m. modulating frequency, a pair of side frequencies is produced. The side frequencies are spaced away from the carrier by an amount equal to the modulation frequency, and their amplitude is directly proportional to the amplitude of the modulation. The amplitude of the carrier does not change under modulation.

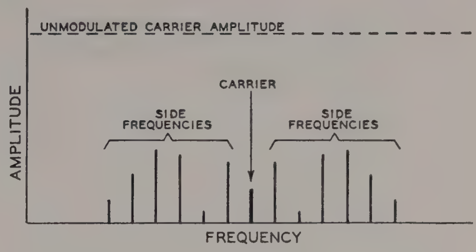


Figure 4.
F.M. SIDE FREQUENCIES.
With f.m. each modulation frequency component causes a large number of side frequencies to be produced. The side frequencies are separated from each other and the carrier by an amount equal to the modulation frequency, but their amplitude varies greatly as the amount of modulation is changed. The carrier strength also varies greatly with frequency modulation. The side frequencies shown represent a case where the deviation each side of the "carrier" frequency is equal to five times the modulating frequency. Other amounts of deviation with the same modulation frequency would cause the relative strengths of the various sidebands to change widely.

ciable frequency modulation of the noise-plus-signal voltage which is applied to the detector in the receiver.

F.M. Terms. Unlike amplitude modulation, the term "percentage modulation" means little in f.m. practice, unless the receiver characteristics are specified. There are, however, three terms, *deviation, modulation index,* and *deviation ratio,* which convey considerable information concerning the character of the f.m. wave.

Deviation is the amount of frequency shift each side of the unmodulated or "resting" carrier frequency which occurs when the transmitter is modulated. Deviation is ordinarily measured in kilocycles, and in a properly operating f.m. transmitter it will be directly proportional to the amplitude of the modulating signal. When a symmetrical modulating signal is applied to the transmitter, equal deviation each side of the resting frequency is obtained during each cycle of the modulating signal, and the total frequency range covered by the f.m. transmitter is sometimes known as the "swing." If, for instance, a transmitter operating on 1000 kc. has its frequency shifted from 1000 kc. to 1010 kc., back to 1000 kc., then to 990 kc., and again back to 1000 kc. during one cycle of the modulating wave, the *deviation* would be 10 kc. and the *swing* 20 kc.

The *modulation index* of an f.m. signal is the ratio of the deviation to the audio modulating frequency, when both are expressed in the same units. Thus, in the example above, if the signal is varied from 1000 kc. to 1010 kc. to 990 kc. and back to 1000 kc. at a rate (frequency) of 2000 times a second, the modulation index would be 5, since the deviation (10 kc.) is 5 times the modulating frequency (2000 cycles, or 2 kc.).

The relative strengths of the f.m. carrier and the various side frequencies depend directly upon the modulation index, these relative strengths varying widely as the modulation index is varied. In the preceding example, for instance, side frequencies occur on the high side of 1000 kc. at 1002, 1004, 1006, 1008, 1010, 1012, etc., and on the low frequency side at 998, 996, 994, 992, 990, 988, etc. In proportion to the unmodulated carrier strength (100 per cent), these side frequencies have the following strengths, as indicated by a modulation index of 5: 1002 and 998—33 per cent, 1004 and 996—5 per cent, 1006 and 994—36 per cent, 1008 and 992—39 per cent, 1010 and 990—26 per cent, 1012 and 988—13 per cent. The carrier strength (1000 kc.) will be 18 per cent of its unmodulated value. Changing the amplitude of the modulating signal will change the deviation, and thus the modulation index will be changed, with the result that the side frequencies, while still located in the same places, will have widely different strength values from those given above.

The *deviation ratio* is similar to the modulation index in that it involves the ratio between a modulating frequency and deviation.

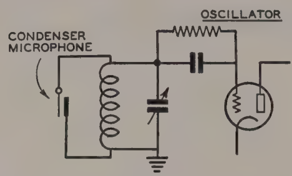


Figure 5.
SIMPLE FREQUENCY MODULATOR.
The variations in capacity of a condenser microphone as sound strikes the diaphragm will cause a corresponding variation in the oscillator frequency.

In this case, however, the deviation in question is the peak frequency shift obtained under full modulation, and the audio frequency to be considered is the maximum audio frequency to be transmitted. When the maximum audio frequency to be transmitted is 5000 cycles, for example, a deviation ratio of 3 would call for a peak deviation of 3 x 5000, or 15 kc. at full modulation. The noise-suppression capabilities of f.m. are directly related to the deviation ratio. As the deviation ratio is increased, the noise suppression becomes better *if* the signal is somewhat stronger than the noise. Where the noise approaches the signal in strength, however, low deviation ratios allow communication to be maintained in many cases where high-deviation-ratio f.m. and conventional a.m. are incapable of giving service. For each value of signal-to-noise ratio at the receiver, there is a maximum deviation ratio which may be used, beyond which the signal becomes smothered in the noise. Up to this critical deviation ratio, however, the noise suppression becomes progressively better as the deviation ratio is increased.

For high-fidelity f.m. broadcasting purposes, a deviation ratio of 5 is ordinarily used, the maximum audio frequency being 15,000 cycles, and the peak deviation at full modulation being 75 kc. Since a swing of 150 kc. is covered by the transmitter, it is obvious that wide-band f.m. transmission must necessarily be confined to the ultra-high frequencies, where room for the signals is available.

For strictly communication work, where the noise-suppression advantages of f.m. must be realized under adverse signal-to-noise ratios, and where maximum coverage for a given amount of power is of prime importance, deviation ratios of 1 to 3 will be found most satisfactory.

Bandwidth Required by F.M. As the above discussion has indicated, many side frequencies are set up when a radio-frequency carrier is frequency modulated; theoretically, in fact, an infinite number of side frequencies is formed. Fortunately, however, the amplitudes of those side frequencies falling outside the frequency range over which the transmitter is "swung" are so small that most of them may be ignored. In f.m. transmission, when a complex modulating wave (speech or music) is used, still additional side frequencies resulting from a beating together of the various frequency components in the modulating wave are formed. This is a situation that does not occur in amplitude modulation and it might be thought that the large number of side frequencies thus formed might make the frequency spectrum produced by an f.m. transmitter prohibitively wide. Analysis shows, however, that the additional side frequencies are of very small amplitude, and, instead of increasing the bandwidth, modulation by a complex wave actually reduces the effective bandwith of the f.m. wave. This is especially true where speech modulation is used, since most of the power in voiced sounds is concentrated at low frequencies around 400 cycles.

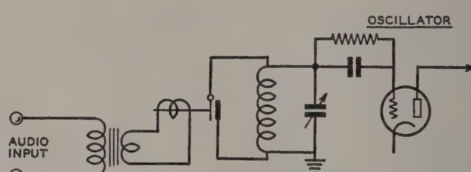


Figure 6.
ELECTRICALLY DRIVEN CONDENSER MODULATOR.
Certain types of audio reproducers, such as earphones and recorders, may be mechanically connected to one plate of a small variable condenser to give frequency modulation. It is important that the driving unit be of the "constant amplitude" type.

When all factors are considered, it is found that an f.m. signal will occupy an effective bandwidth of about 2½ times the maximum swing at peak modulation.

Modulating Circuits

A successful frequency modulated transmitter must meet two requirements: (1) The frequency deviation must be symmetrical about a fixed frequency, for symmetrical modulation voltage. (2) The deviation must be directly proportional to the amplitude of the modulation, and independent of the modulation frequency. There are several methods of fre-

quency modulation which will fulfill these requirements. Some of these methods will be described in the following paragraphs.

Mechanical Modulators. The arrangement shown in Figure 5 is undoubtedly the simplest of all frequency modulators. A condenser microphone is connected across the oscillator tank circuit, and the variations in capacity produced by the microphone cause the oscillator frequency to vary at the frequency of the impressed sound. Since condenser microphones are difficult to obtain, and the amount of r.f. voltage which may be safely impressed across them is small, the circuit is of little practical use, however. Figure 6 shows a modification of Figure 5 which is more suited to practical application. Here the variable capacity device which varies the frequency consists of a condenser, one plate of which is

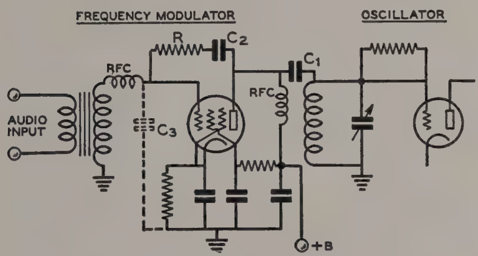


Figure 7.
REACTANCE-TUBE MODULATOR.
This is a popular form of frequency modulator. The operation of the circuit is described in the text. Typical values for the components will be found in similar circuits shown in Chapter 19.

moved by being mechanically coupled to an electro-mechanical driving unit such as a loud speaker or phonograph recording head. This circuit, while practical, is seldom used, because most driving units do not give frequency modulation which complies with requirement (2). The requirement is met by piezo-electric (crystal) reproducers such as earphones and recorders, however, and this type of "constant amplitude" driving unit may be used successfully.

Reactance-Tube Modulators. One of the most practical ways of obtaining frequency modulation is through the use of a *reactance-tube modulator.* In this arrangement the modulator plate-cathode circuit is connected across the oscillator tank circuit, and made to appear as either a capacitive or inductive reactance by exciting the modulator grid with a voltage which either leads or lags the oscillator tank voltage by 90 degrees. The leading or lagging grid voltage causes a corresponding leading or lagging plate current, and the plate-cathode circuit appears as a capacitive or inductive reactance across the oscillator tank circuit. When the transconductance of the modulator tube is varied, by varying one of the element voltages, the magnitude of the reactance across the oscillator tank is varied. By applying audio modulating voltage to one of the elements, the transconductance, and hence the frequency, may be varied at an audio rate. When properly designed and operated, the reactance-tube modulator gives linear frequency modulation, and is capable of producing large amounts of deviation.

There are numerous possible configurations of the reactance-tube modulator circuit. The difference in the various arrangements lies principally in the type of phase-shifting circuit used to give a grid voltage which is in phase quadrature with the r.f. voltage at the modulator plate.

Figure 7 is a diagram of one of the most popular forms of reactance-tube modulators. The modulator tube, which is usually a sharp cutoff pentode such as a 6J7 or 6SJ7, has its plate coupled through a blocking condenser, C_1, to the "hot" side of the oscillator grid circuit. Another blocking condenser, C_2, feeds r.f. to the phase shifting network R-C_3 in the modulator grid circuit. If the resistance of R is made large in comparison with the reactance of C_3 at the oscillator frequency, the current through the R-C_3 combination will be nearly in phase with the voltage across the tank circuit, and the voltage across C_3 will lag the oscillator tank voltage by almost 90 degrees. The result of the 90-degree lagging voltage on the modulator grid is that its plate current lags the tank voltage by 90 degrees, and the react-

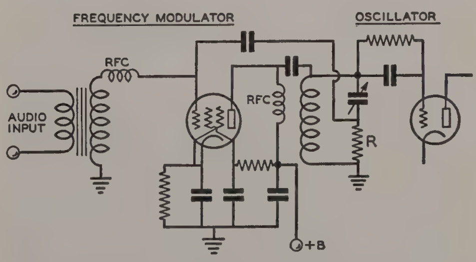


Figure 8.
REACTANCE TUBE MODULATOR.
This circuit operates similarly to the one shown in Figure 7. The difference between the two lies in the method in which the r.f. grid voltage is shifted 90 degrees in phase from the r.f. plate voltage.

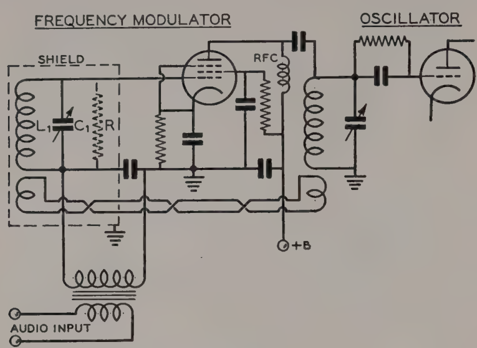


Figure 9.
TUNED PHASE-SHIFT CIRCUIT.
By using a tuned circuit, L_1-C_1, to shift the phase of the reactance-tube grid excitation, the phase shift may be adjusted to reduce the loading on the oscillator under modulation.

ance tube appears as an inductance in shunt with the oscillator inductance, thus raising the oscillator frequency.

The phase-shifting condenser C_3 is usually provided by the input capacitance of the modulator tube and stray capacity between grid and ground, and it will not ordinarily be found necessary to employ an actual condenser for this purpose at frequencies above 2 or 3 Mc. Resistance R will usually have a value of between 25,000 and 100,000 ohms. Either resistance or transformer coupling, as shown, may be used to feed audio voltage to the modulator grid. When a resistance coupling is used, it is necessary to shield the grid circuit adequately, since the high impedance grid circuit is prone to pick up stray r.f. and low frequency a.c. voltage, and cause undesired frequency modulation. Another disadvantage to the use of a resistance in the grid circuit is that small amounts of grid current may bias the grid of the reactance tube to the point where its effectiveness as a modulator is reduced considerably.

Another of the numerous practical reactance-tube circuits is shown in Figure 8. In this circuit, the 90-degree phase shift in grid excitation to the modulator is obtained by placing a resistor in series with the oscillator tank condenser. Since the current through the tank condenser leads the voltage across the tank circuit by 90 degrees, the r.f. voltage applied to the modulator grid will also lead this voltage by the same amount; the modulator plate current will lead the tank voltage, and the modulator tube will appear as a condenser.

The resistor, R, may be placed in series with the tank coil, rather than the condenser, in which case the phase relationships are such that the reactance tube appears as an inductance. Too much resistance in either leg of the oscillator tank will result in such a low Q circuit that it will be impossible to maintain oscillation. Carbon resistors of around 25 ohms will provide sufficient excitation to the modulator for good sensitivity.

There are several possible variations of the basic reactance-tube modulator circuits shown in Figures 7 and 8. The audio input may be applied to the suppressor grid, rather than the control grid, if desired. This method requires that the control grid be returned to ground through a rather high resistance (250,000 ohms to 1 megohm) or through an r.f. choke. Another modification is to apply the audio to a grid other than the control grid in a mixer or pentagrid converter tube which is used as the modulator. Generally, it will be found that the transconductance variation per volt of control-element voltage variation will be greatest when the control (audio) voltage is applied to the control grid. In cases where it is desirable to separate completely the audio and r.f. circuits, however, applying audio voltage to one of the other elements will often be found advantageous in spite of the somewhat lower sensitivity.

In spite of the fact that high-plate resistance pentodes are usually used as reactance tubes, it will often be found that amplitude modulation due to loading of the oscillator by the reactance tube takes place when a large amount of frequency modulation is attempted. The cure for this type of amplitude modulation will usually be found in adjusting the phase of the r.f. voltage applied to the reactance tube grid until it differs somewhat from the recommended 90-degree relation with the r.f. at the plate. One such method consists of using the reactance-tube circuit shown in Figure 7 in conjunction with a Hartley or Colpitts oscillator, in which the center of the oscillator tank circuit is grounded for r.f. In this case, both ends of the oscillator coil will be equally "hot," and the C_2-R combination may be connected to the opposite end of the tank circuit from which the reactance-tube plate is connected. Then, by adjustment of C_3 or R, the phase shift between grid and plate may be made more than 90 degrees, and amplitude modulation balanced out.

A circuit which allows the phase shift to be set exactly at 90 degrees, or to be varied either way, is shown in Figure 9. This circuit uses a separate tuned circuit in the reactance-tube grid. The additional circuit may be coupled to the oscillator either by a link, as shown, or simply by placing the two coils close to each other. When the L_1-C_1 circuit is tuned to resonance, the voltage developed across it will

be 90 degrees out of phase with the voltage across the oscillator tank. Detuning the L_1-C_1 circuit in one direction or the other will cause the phase shift to become greater or less than 90 degrees.

To reduce the excitation applied to the grid of the reactance tube and to make the tuning of the phase-shifting network less critical, a resistance, R, may be placed across the circuit. The resistor may have a value as low as a few hundred ohms, and it will be found that large changes in the value of resistance will make it necessary to change the setting of C_1 to maintain the correct amount of phase shift.

Adjusting the Phase Shift. One of the simplest methods of adjusting the phase shift to the correct amount is to place a pair of earphones in series with the oscillator cathode-to-ground circuit and adjust the phase-shift network until minimum sound is heard in the 'phones when frequency modulation is taking place. If an electron-coupled or Hartley oscillator is used, this method requires that the cathode circuit of the oscillator be inductively or capacitively coupled to the grid circuit, rather than tapped on the grid coil. The 'phones should be adequately by-passed for r.f., of course.

Stabilization. Due to the presence of the frequency modulator, the stabilization of an f.m. oscillator in regard to voltage changes is considerably more involved than in the case of a simple self-controlled oscillator for transmitter frequency control. If desired, the oscillator itself may be made perfectly stable under voltage changes, but the presence of the frequency modulator destroys the beneficial effect of any such stabilization. It thus becomes desirable to apply the stabilizing arrangement to the modulator as well as the oscillator. If the oscillator itself is stable under voltage changes, or, in other words, self-compensated by some means such as the use of an electron-coupled circuit, it is only necessary to apply voltage-frequency compensation to the modulator. Stabilized power supply arrangements suitable for use on the modulator or both modulator and oscillator are described fully in Chapter 15.

A circuit in which automatic stabilization of the effects of voltage variations on the modulator is obtained, is shown in Figure 10. In this circuit, the reactance-tube grids are connected in push-pull across the phase-shifting circuit L_1-C_1, while the plates are connected in parallel and tied to the oscillator tank in the usual manner. Any variation in the plate-supply voltage to the reactance tubes causes equal and opposite effects in their reactance, and there is no net reactance variation.

Another method of oscillator stabilization makes use of a discriminator circuit. This arrangement stabilizes the frequency against changes arising from any cause (except the

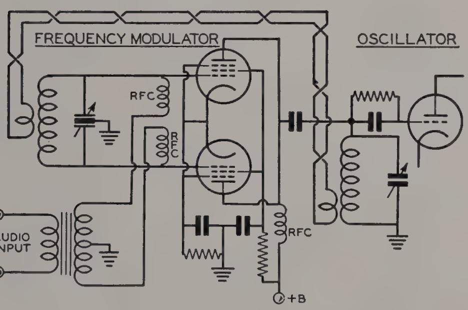


Figure 10.
STABILIZED REACTANCE-TUBE MODULATOR.

Frequency shift due to voltage changes on the modulator may be greatly reduced by the use of this circuit. Changes in element voltages cause equal and opposite changes in reactance in the two modulators, thus minimizing the frequency shift. The reactance-tubes grids receive excitation from a balanced tuned circuit so that one tube receives voltage lagging the oscillator tank voltage by 90°, while the other tube is excited with a voltage that leads the tank voltage by 90°.

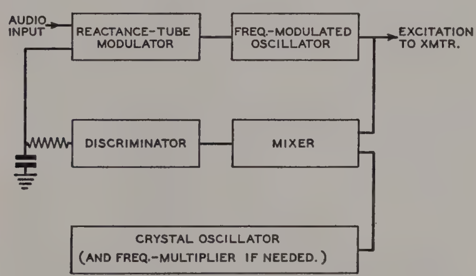


Figure 11.
DISCRIMINATOR STABILIZING ARRANGEMENT.

A frequency-modulated oscillator may be stabilized against undesired frequency shift by comparing the transmitter frequency with that of a crystal oscillator. The difference between the two frequencies is applied to a discriminator circuit, and any change from a predetermined difference will cause the discriminator to restore the transmitter to its correct frequency. An R-C filter is used to remove the audio modulation from the discriminator output.

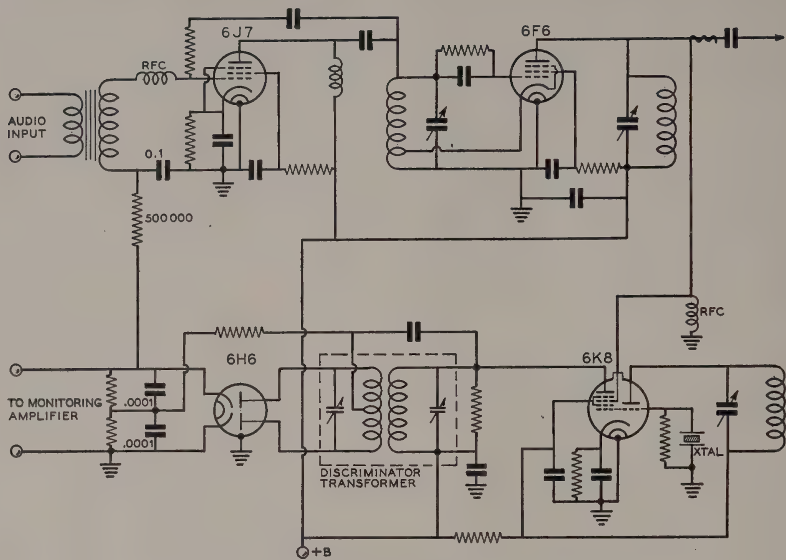


Figure 12.
TYPICAL STABILIZATION CIRCUIT.
A schematic diagram of the arrangement shown in block-diagram style in Figure 11. For maximum sensitivity, the discriminator should operate on a frequency around 455 kilocycles.

desired modulation) by comparing the oscillator frequency with a crystal controlled standard and applying the proper compensating voltages. A block diagram of this method is shown in Figure 11. Output from one of the stages of the transmitter is mixed with the output of a crystal oscillator to give an "intermediate frequency" output which is applied to a conventional discriminator. The discriminator, which will be more completely described later in this chapter, is a circuit arrangement to produce an output voltage which depends on the frequency of the r.f. applied to it.

The d.c. voltage produced by the discriminator is applied to a reactance tube tied across the oscillator tank circuit. As the average or "center" frequency varies one way or the other from the correct value, a positive or negative voltage appears across the discriminator load resistors. When this voltage is placed on the control element of the reactance tube, it attempts to restore the center (mid-modulation or unmodulated) radio frequency to a value which gives zero voltage output from the discriminator. The oscillator can never be fully restored to its correct frequency, however, since the discriminator output voltage would then be zero, and no frequency correction would be taking place. The frequency is actually shifted back to a value somewhere between what it should be and what it would have been without stabilization. The reactance tube which takes care of the frequency correction may also be used as the modulator, and the frequency stabilizing voltage may be applied in series with the audio voltage or, alternatively, it may be applied to another of the tube elements. The audio output of the discriminator must be removed by a simple R-C filter so that the compensating voltage is direct current without superimposed audio. The audio output of the discriminator may be used for monitoring purposes, if desired. Obviously the stability of the complete arrangement is dependent upon the stability of the discriminator components under temperature and humidity changes, and upon the stability of the crystal oscillator. Ordinarily the stability of the crystal oscillator will be sufficiently great that the discriminator will be the limiting factor in the amount of stabilization obtainable, making it necessary to use discriminator components (especially the tuned input transformer) of good quality. A typical stabilizing circuit, with provision for monitoring, is shown in Figure 12.

The frequency of the crystal used in the stabilizing circuit will depend upon the fre-

quency at which the discriminator operates, and the frequency of the stage in the transmitter from which the stabilizer signal is taken. If a b.c. replacement type discriminator transformer designed for a frequency in the 400–500 kc. range is used, the r.f. input for the stabilizer may be obtained from the transmitter oscillator stage, or, if more sensitivity is desired, from the plate circuit of the frequency multiplier following the oscillator. The crystal oscillator must operate on a frequency such that its fundamental, or one of its harmonics, falls at a frequency which differs from that of the transmitter stage applied to the stabilizer by an amount equal to the discriminator frequency. If the required crystal frequency falls higher than is easily obtainable with a crystal, it may be necessary to use a frequency multiplier after the crystal stage.

Linearity Test. It is almost a necessity to run a static test on the reactance-tube frequency modulator to determine its linearity and effectiveness, since small changes in the values of components, and in stray capacities, will almost certainly alter the modulator characteristics. A frequency-versus-control-voltage curve should be plotted to ascertain that equal increments in control voltage, both in a positive and a negative direction, cause equal changes in frequency. If the curve shows that the modulator has an appreciable amount of non-linearity, changes in bias, electrode voltages, and resistance values may be made to obtain a straight-line characteristic.

Figure 13 shows a method of connecting two $4\frac{1}{2}$-volt C batteries and a potentiometer to plot the characteristic of the modulator. It will be necessary to use a zero-center voltmeter to measure the grid voltage, or else reverse the voltmeter leads when changing from positive to negative grid voltage. When a straight-line characteristic for the modulator is obtained by the static test method, the capacities of the various by-pass condensers in the circuit must be kept small to retain this characteristic when an audio voltage is used to vary the frequency in place of the d.c. voltage with which the characteristic was plotted.

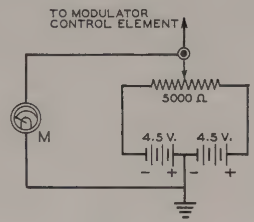


Figure 13.

This circuit allows the control characteristic of the frequency modulator to be easily checked. As the potentiometer arm is moved one way or the other from the center position, a positive or negative voltage is placed on the modulator control element.

Phase Modulation—the Armstrong System. By means of phase modulation (p.m.) it is possible to dispense with self-controlled oscillators, and obtain directly crystal-controlled f.m. In the final analysis, p.m. is simply frequency modulation in which the deviation is directly proportional to the modulation frequency. If an audio modulating signal of 1000 cycles causes a deviation of $\frac{1}{2}$ kc., for example, a 2000-cycle modulating signal of the same amplitude will give a deviation of 1 kc., and so on. To produce an f.m. signal, it is necessary to make the deviation independent of the modulation frequency, and proportional only to the amplitude of the modulating signal. With p.m., this is done by including a frequency correcting network in the audio system of the transmitter. The audio correction network must have an attenuation that varies directly with frequency, and this requirement is easily met by a very simple resistance-capacity network.

While the equipment required for a p.m. type of f.m. transmitter is rather complex, its operation and adjustment are quite simple. In fact, it is a question whether the p.m. method will not ultimately prove more satisfactory for low-deviation communication-type f.m. than the reactance-tube method, when the necessary frequency-controlling apparatus required by the latter system is considered. Actually, the circuits for obtaining p.m. are no more complicated than the reactance-tube f.m. circuits. The complications with p.m. arise from the fact that very little actual frequency modulation is produced by the phase-modulating method, and a great deal of frequency multiplication is necessary to obtain a reasonable amount of deviation. The frequency multiplication may be carried out at very low power levels, however, with low-cost, small receiving parts.

Odd-harmonic distortion is produced when f.m. is obtained by the phase-modulation method, and the amount of this distortion that can be tolerated is the limiting factor in determining the amount of p.m. that can be used. Since the aforementioned frequency-correcting network causes the lowest modulating frequency to have the greatest amplitude, maximum phase modulation takes place at the lowest modulating frequency, and the

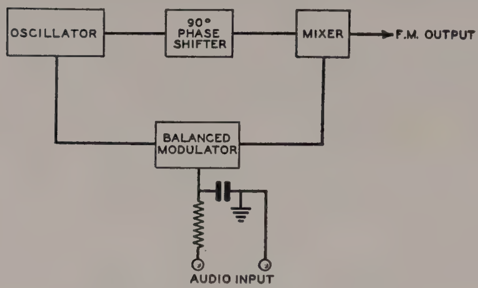


Figure 14.
PHASE MODULATOR BLOCK DIAGRAM.

The R-C network in the audio input leads makes the amount of phase modulation inversely proportional to the audio frequency, thus giving frequency modulated output.

amount of distortion that can be tolerated at this frequency determines the maximum deviation that can be obtained by the p.m. method. For high-fidelity broadcasting, the deviation produced by p.m. is limited to an amount equal to about one-third of the lowest modulating frequency. When the lowest modulating frequency is 30 cycles, the deviation is thus limited to about 10 cycles, and it may be seen that an enormous amount of frequency multiplication is necessary to get a deviation of, say, 75 kc. on the output frequency.

In voice modulation, the peak intensity occurs at around 400 cycles and, with a slight increase in distortion, it is possible to increase the deviation until it equals one-half this frequency. The deviation would thus be 200 cycles, and the necessary amount of frequency multiplication can be greatly reduced. With a constant deviation of 200 cycles, the distortion produced at a modulation frequency of 400 cycles is about 7 per cent, with higher distortion at lower frequencies and negligible distortion at higher frequencies. Fortunately, the distortion at frequencies lower than 400 cycles is not of great importance, since the amplitude of these components is low enough so that the amount of modulation deviation is reduced, with a corresponding decrease in the actual distortion.

P.M. Circuits. To obtain phase modulation it is only necessary to shift the phase of the sidebands produced in amplitude modulation by 90 degrees. When the phase-shifted sidebands are re-combined with the carrier, the result is a phase-modulated signal. A block diagram of the basic p.m. arrangement is shown in Figure 14. Output from a crystal or other high-stability oscillator is supplied to two circuit branches. In one of the branches is a balanced modulator, which produces sidebands, minus the carrier, when modulation is applied. The other branch, which contains a mixer, is fed through a phase-shifting circuit. The output of the two branches is combined in the plate circuit of the mixer, and the result is a phase-modulated carrier. By feeding the audio to the modulator through the R-C attenuator network shown, the phase modulation is made inversely proportional to frequency, and the final result is a frequency-modulated signal. Although the diagram shows the phase-shifting network between the oscillator and mixer tubes, it does not necessarily have to be in this position; quite often it will be found in either the input or output circuits of the balanced-modulator stage. The only requirement is that there be a 90-degree phase difference between the two components supplied to the mixer, and the location of the phase-shift network is largely a matter of convenience.

To increase the small amount of deviation produced by the p.m. method to a useable amount, a considerable amount of frequency multiplication is needed between the exciter and the transmitter output stage. The necessary frequency multiplication may be obtained in either one of two ways: The oscillator frequency may be made low enough so that the frequency multiplication necessary to reach the desired output frequency will be sufficient to give the desired deviation, or a moderately high frequency oscillator may be followed by a small amount of frequency multiplication, and the signal then "beaten back" by means of a heterodyne oscillator and a mixer to another moderately high-frequency, whence it may be multiplied in the usual manner to the output frequency. As an example of the first method, a crystal oscillator using a 460- to 465-kc. filter type crystal may have its frequency increased 128 times by a string of doublers or quadruplers to reach the 58.5- to 60-Mc. amateur f.m. band. If the original maximum deviation was 200 cycles, the deviation on the output frequency would thus be 25.6 kc.

An example of the second method is the use of a crystal oscillator, followed by the phase modulator, on 1800 kc. The p.m. output is tripled to 5400 kc., where the deviation is then 3 times what it originally was. Beating the 5400-kc. output with another crystal oscillator on 7250 kc. gives a difference of frequency of 1850 kc., with the deviation still tripled from its original value. By a series of doublers or quadruplers the 1850-kc. signal may be multiplied 32 times to reach a frequency of 59.2 Mc., which is also in the 58.5- to 60-Mc. amateur band. The increase in deviation will be

equal to the product of the two frequency multiplications (3 x 32) or 96 times. A p.m. exciter using this type of frequency multiplication is described in Chapter 19, while a block diagram of the basic method is shown in Figure 15.

Measurement of Deviation. When a single-frequency modulating voltage is used with an f.m. transmitter, the relative amplitudes of the various sidebands and the carrier vary widely as the deviation is varied by increasing or decreasing the amount of modulation. Since the relationship between the amplitudes of the various sidebands and carrier to the audio modulating frequency and the deviation is known, a simple method of measuring the deviation of a frequency modulated transmitter is possible. In making the measurement, the result is given in the form of the modulation index for a certain amount of audio input. As previously described, the modulation index is the ratio of the peak frequency deviation to the frequency of the audio modulation.

The measurement is made by applying a sine-wave audio voltage of known frequency to the transmitter, and increasing the modulation until the amplitude of the carrier component of the frequency modulated wave reaches zero. The modulation index for zero carrier may then be determined from the table below. As may be seen from the table, the first point of zero carrier is obtained when the modulation index has a value of 2.405,—in other words, when the deviation is 2.405 times the modulation frequency. For example, if a modulation frequency of 1000 cycles is used, and the modulation is increased until the first carrier null is obtained, the deviation will then be 2.405 times the modulation frequency, or 2.405 kc. If the modulating frequency happened to be 2000 cycles, the deviation at the first null would be 4.810 kc. Other carrier nulls will be obtained when the index is 5.52, 8.654, and at increasing values separated approximately by π. The following is a listing of the modulation index at successive carrier nulls up to the tenth:

Zero carrier point no.	Modulation index
1	2.405
2	5.520
3	8.654
4	11.792
5	14.931
6	18.071
7	21.212
8	24.353
9	27.494
10	30.635

The only equipment required for making the measurements is a calibrated audio oscillator of good wave form, and a communication receiver equipped with a beat oscillator and crystal filter. The receiver should be used with its crystal filter set for a bandwidth of approximately twice the modulation frequency, to exclude sidebands spaced from the carrier by the modulation frequency. The unmodulated carrier is accurately tuned in on the receiver with the beat oscillator operating, and modulation from the audio oscillator is then applied to the transmitter, and the modulation increased until the first carrier null is obtained. This first carrier null will correspond to a modulation index of 2.405, as previously mentioned. Successive null points will correspond to the indices listed in the table. A volume indicator in the transmitter audio system may be used to measure the audio level required for different amounts of deviation, and the indicator thus calibrated in terms of frequency deviation. If the measurements are made at the fundamental frequency of the oscillator,

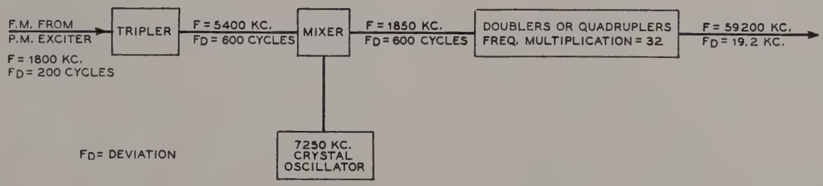


Figure 15.
P.M. DEVIATION-INCREASING SCHEME.
The small amount of f.m. deviation caused by phase modulation may be increased to useful amounts by multiplying the frequency somewhat, heterodyning it down to a low frequency, and again multiplying to the output frequency. The increase in deviation is equal to the product of the separate frequency multiplications, or, in the case shown above 3 × 32, or 96 times.

it will be necessary to multiply the frequency deviation by the harmonic upon which the transmitter is operating, of course. It will probably be most convenient to make the determination at some frequency intermediate between that of the oscillator and that at which the transmitter is operating, and then multiply the result by the frequency multiplication between that point and the transmitter output frequency.

Frequency-Modulation Reception

In contrast with the transmitter, where the use of f.m. greatly simplifies the modulation problem, for serious work the use of f.m. necessitates a receiver somewhat more complicated than would be necessary for amplitude modulation. While ordinary superheterodyne, t.r.f., and superregenerative receivers will receive f.m. after a fashion, serious work requires a receiver especially designed for f.m. reception.

The f.m. receiver must have, first of all, a bandwidth sufficient to pass the range of frequencies generated by the f.m. transmitter. And since the receiver must be a superheterodyne if it is to have good sensitivity at the frequencies to which f.m. is restricted, i.f. bandwidth is an important factor in its design.

The second requirement of the f.m. receiver is that it incorporate some sort of device for converting frequency changes into amplitude changes, in other words, a detector operating on frequency variations rather than amplitude variations. The third requirement, and one which is necessary if the full noise reducing capabilities of the f.m. system of transmission are desired, is a limiting device to eliminate amplitude variations before they reach the detector. A block diagram of the essential parts of an f.m. receiver is shown in Figure 16.

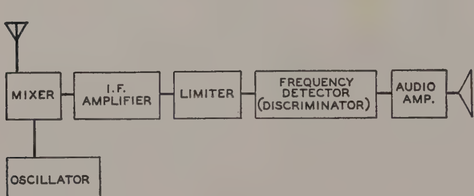


Figure 16.
RECEIVER BLOCK DIAGRAM.
Up to the amplitude limiter stage, the f.m. receiver is similar to an a.m. receiver, except for a somewhat wider i.f. bandwidth. The limiter removes any amplitude modulation, and the frequency detector following the limiter converts frequency variations into amplitude variations.

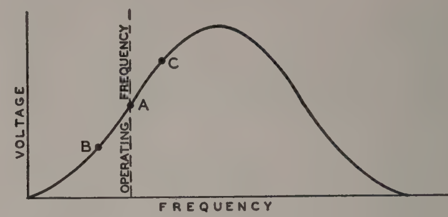


Figure 17.
"OFF TUNE" FREQUENCY DETECTOR.
A portion of the resonance characteristic of a tuned circuit may be used as shown to convert frequency variations into amplitude variations.

The Frequency Detector. The simplest device for converting frequency variations to amplitude variations is an "off-tune" resonant circuit, as illustrated in Figure 17. With the carrier tuned in at point "A," a certain amount of r.f. voltage will be developed across the tuned circuit, and, as the frequency is varied either side of this frequency by the modulation, the r.f. voltage will increase and decrease to points "C" and "B" in accordance with the modulation. If the voltage across the tuned circuit is applied to an ordinary detector, the detector output will vary in accordance with the modulation, the amplitude of the variation being proportional to the deviation of the signal, and the rate being equal to the modulation frequency. It is obvious from Figure 17 that only a small portion of the resonance curve is usable for linear conversion of frequency variations into amplitude variations, since the linear portion of the curve is rather short. Any frequency variation which exceeds the linear portion will cause distortion of the recovered audio.

Travis Discriminator. Another form of frequency detector or *discriminator,* is shown in Figure 18. In this arrangement two tuned circuits are used, one tuned on each side of the i.f. amplifier frequency, and with their resonant frequencies spaced slightly more than the expected transmitter "swing" apart. Their outputs are combined in a differential rectifier so that the voltage across the series load resistors, R_1 and R_2, is equal to the algebraic sum of the individual output voltages of each rectifier. When a signal at the i.f. mid-frequency is received, the voltages across the load resistors are equal and opposite, and the sum voltage is zero. As the r.f. signal varies from the mid-frequency, however, these individual voltages become unequal, and a voltage having the polarity of the largest voltage and

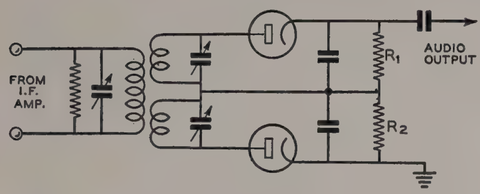


Figure 18.
TRAVIS DISCRIMINATOR.

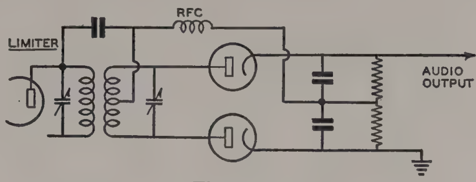


Figure 20.
FOSTER-SEELEY DISCRIMINATOR.
This discriminator depends on the phase relationships between a primary and a tuned secondary for its operation.

equal to the difference between the two voltages appears across the series resistors, and is applied to the audio amplifier. The relationship between frequency and discriminator output voltage is shown in Figure 19. The separation of the discriminator peaks and the linearity of the output voltage vs. frequency curve depend upon the discriminator frequency, the Q of the tuned circuits, and the value of the diode load resistors. As the intermediate (and discriminator) frequency is increased, the peaks must be separated farther to secure good linearity and output. Within limits, as the diode load resistors or the Q are reduced, the linearity improves, and the separation between the peaks must be greater.

Foster-Seeley Discriminator. The most widely used form of discriminator is that shown in Figure 20. This type of discriminator yields an output-voltage-versus-frequency characteristic similar to that shown in Figure 19. Here, again, the output voltage is equal to the algebraic sum of the voltages developed across the load resistors of the two diodes, the resistors being connected in series to ground. However, this *Foster-Seeley* discriminator requires only two tuned circuits instead of the three used in the previous discriminator. The operation of the circuit results from the phase relationships existing in a transformer having a tuned secondary. In effect, as a close examination of the circuit will reveal, the primary circuit is in series, for r.f., with each half of the secondary to ground. When the received signal is at the resonant frequency of the secondary, the r.f. voltage across the secondary is 90 degrees out of phase with that across the primary. Since each diode is connected across one half of the secondary winding and the primary winding in series, the resultant r.f. voltages applied to each are equal, and the voltages developed across each diode load resistor are equal and of opposite polarity. Hence, the net voltage between the top of the load resistors and ground is zero. This is shown vectorially in Figure 21A, where the resultant voltages R and R′ which are applied

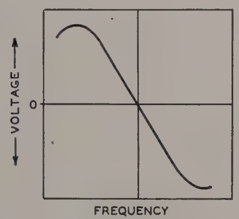


Figure 19.
DISCRIMINATOR VOLTAGE-FREQUENCY CURVE.
At its "center" frequency the discriminator produces zero output voltage. On either side of this frequency it gives a voltage of a polarity and magnitude which depend on the direction and amount of frequency shift.

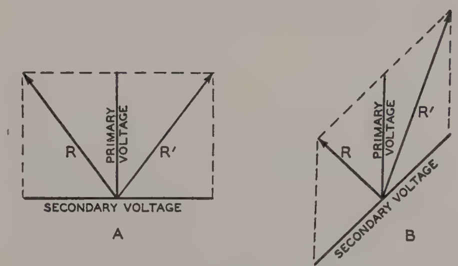


Figure 21.
DISCRIMINATOR VECTOR DIAGRAM.
A signal at the resonant frequency of the secondary will cause the secondary voltage to be 90 degrees out of phase with the primary voltage, as shown at A, and the resultant voltages R and R′ are equal. If the signal frequency changes, the phase relationship also changes, and the resultant voltages are no longer equal, as shown at B. A differential rectifier is used to give an output voltage proportional to the difference between R and R′.

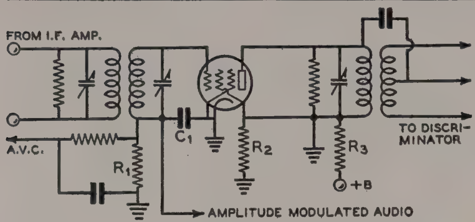


Figure 22.
LIMITER CIRCUIT.

The limiter stage overloads easily, and when overloaded will not reproduce amplitude variations. R_1 may have a value of from 250,000 ohms to 1 megohm. Condenser C_1 should be rather small, about .0001 μfd. Resistors R_2 and R_3 should be proportioned so that the plate and screen voltage is from 10 to 30 volts.

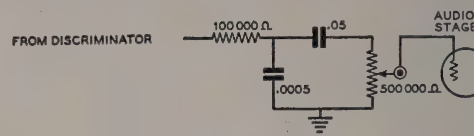


Figure 23.
LOW-PASS FILTER.

A low-pass filter is necessary in the f.m. receiver to remove high frequency noise components.

to the two diodes are shown to be equal when the phase angle between primary and secondary voltages is 90 degrees. If, however, the signal varies from the resonant frequency, the 90-degree phase relationship no longer exists between primary and secondary. The result of this effect is shown in Figure 21B, where the secondary r.f. voltage is no longer 90 degrees out of phase with respect to the primary voltage. The resultant voltages applied to the two diodes are now no longer equal, and a d.c. voltage proportional to the difference between the r.f. voltages applied to the two diodes will exist across the series load resistors. As the signal frequency varies back and forth across the resonant frequency of the discriminator, an a.c. voltage of the same frequency as the original modulation, and proportional to the deviation, is developed and passed on to the audio amplifier.

Limiters. The limiter in an f.m. receiver serves to remove amplitude modulation and pass on to the discriminator a frequency modulated signal of constant amplitude; a typical circuit is shown in Figure 22. The limiter tube is operated as an i.f. stage with very low plate voltage and with grid leak bias, so that it overloads quite easily. Up to a certain point the output of the limiter will increase with an increase in signal. Above this point, however, the limiter becomes overloaded, and further large increases in signal will not give any increase in output. To operate successfully, the limiter must be supplied with a large amount of signal, so that the amplitude of its output will not change for rather wide variations in amplitude of the signal. Noise, which causes little frequency modulation but much amplitude modulation of the received signal is virtually wiped out in the limiter.

The voltage across the grid resistor, R_1, varies with the amplitude of the received signal, and for this reason, conventional amplitude modulated signals may be received on the f.m. receiver by connecting the input of the audio amplifier to the top of this resistor, rather than to the discriminator output. When properly filtered by a simple R-C circuit, the voltage across R_1 may also be used as a.v.c. voltage for the receiver. When the limiter is operating properly, a.v.c. is neither necessary nor desirable, however.

Receiver Design Considerations. One of the most important factors in the design of an f.m. receiver is the frequency swing which it is intended to handle. It will be apparent from Figure 19 that if the straight portion of the discriminator circuit covers a wider range of frequencies than those generated by the transmitter, the audio output will be reduced from the maximum value of which the receiver is capable.

In this respect, the term "modulation percentage" is more applicable to the f.m. receiver than it is to the transmitter, since the "modulation capability" of the communication system is limited by the receiver bandwidth and the discriminator characteristic; full utilization of the linear portion of the characteristic amounts, in effect, to "100 per cent" modulation. This means that some sort of standard should be agreed upon, for any particular type of communication, to make it unnecessary to vary the transmitter swing to accommodate different receivers.

Two considerations influence the receiver bandwidth necessary for any particular type of communication. These are the maximum audio frequency which the system will handle, and the deviation ratio which will be employed. For voice communication, the maximum audio frequency is more or less fixed at 3000 to 4000 cycles. In the matter of deviation ratio, however, the amount of noise suppression which the f.m. system will provide is influenced by the ratio chosen, since the improvement in signal-to-noise ratio which the f.m. system shows over amplitude modulation is equiva-

lent to a constant *multiplied by the deviation ratio.* This assumes that the signal is somewhat stronger than the noise at the receiver, however, as the advantages of wide-band f.m. in regard to noise suppression disappear when the signal-to-noise ratio approaches unity.

On the other hand, a low deviation ratio is likely to be more satisfactory for strictly communication work, where readability at low signal-to-noise ratios is more important than additional noise suppression when the signal is already appreciably stronger than the noise.

As mentioned previously, broadcast f.m. practice is to use a deviation ratio of 5. When this ratio is applied to a voice-communication system, the total "swing" becomes 30 to 40 kc. With lower deviation ratios, such as are sometimes used for voice work, the swing becomes proportionately less, until at a deviation ratio of 1 the swing is equal to twice the highest audio frequency. Actually, however, the receiver bandwidth must be slightly greater than the expected transmitter swing, since for distortionless reception the receiver must pass the complete band of energy generated by the transmitter, and this band will always cover a range somewhat wider than the transmitter swing.

No definite recommendations can be given for the proper reception bandwidth to be used in a receiver designed for voice f.m. work, since there is as yet no standardization in regard to deviation ratio and maximum audio frequency. For best results, however, it should be remembered that the receiver must have a bandwidth sufficiently wide to pass all of the side frequencies of appreciable strength produced by the transmitter. At the same time, the selectivity of the i.f. channel should be such that the pass band is no wider than is absolutely necessary, since the additional bandwidth in the receiver only serves to decrease the signal-to-noise ratio.

Audio Bandwidth. To realize the full noise reducing capabilities of f.m., it is essential that the pass band of the audio section of the receiver be limited to that necessary for communication. The noise output of the discriminator is proportional to the audio frequency of the noise, and the improvement in signal-to-noise ratio depends almost entirely on *receiver* deviation ratio, or the ratio between one-half the r.f. bandwidth and the audio bandwidth.

A suitable arrangement for removing frequencies higher than those necessary for communication is shown in Figure 23. The 100,000-ohm resistor and the .0005-μfd. condenser reduce the high frequency audio which is passed to the audio amplifier.

CHAPTER TEN

Transmitting Tubes

In THE following table, tubes for low- and medium-power transmitting applications are shown by manufacturer's type number in numerical order, regardless of prefix or suffix letters.

Certain of the tubes listed (800, 830-B, 865, etc.) are manufactured primarily for replacement of such tubes already in use, as they have been superseded by improved models which cost less than their prototypes. Therefore, when choosing tubes for a transmitter which is being designed, it is wise to study the characteristics and prices of all tubes in the same class before making a decision. Flat plate tubes, which necessarily have comparatively high interelectrode capacities (810, etc.), generally are easier to drive than other tubes at low frequencies, but do not work as well at u.h.f. as do "low C" tubes having cylindrical plates (35T, TW-75, HK-54, 808, etc.). However, nearly all types operate satisfactorily on frequencies below 7.0 Mc.

Socket connection diagrams and footnote references will be found on page 237.

TRANSMITTING TUBES

2 Watts	HY24, HY114B
2.5 Watts	RK33
3 Watts	HY63
3.5 Watts	HY6J5GTX, HY615
5 Watts	1626
6 Watts	RK24, 1610
10 Watts	RK23, RK25, RK25B, RK34, RK45, HY65, 802, 1613
12 Watts	RK44, 837
15 Watts	RK10, HK14, HY60, HY75, RK100, WE307A, 832, 841 (RCA), 843, 844, 865, HY6V6GTX, 1619
20 Watts	T20, TZ20, 801, 801A, 1608, 310
21 Watts	T21, HY6L6GX, RK49, 1614
25 Watts	RK11, RK12, HK24, HY25, RK28, RK30, RK39, RK41, HY61, WE254B, 1624
30 Watts	HY30Z, HY31Z, WE316A, 807*, 809*, HY1231Z, 1623*, 1625*
35 Watts	800
40 Watts	RK18, RK20A, RK31, HY40, HY40Z, T40, TZ40, RK46, HY69, WE300A, 804, 829, HY1269, 1628
50 Watts	RK32, RK35, RK37, RK47, HK54, WE304B, 808, 834, 841SW (Taylor)
55 Watts	T55, 811*, 812*
60 Watts	RK51, WE305A, 830B, 826
62.5 Watts	RK52
65 Watts	HY51A, HY51B, HY51Z, 203Z, 814* (RCA-G.E.), HY67
70 Watts	35T, WE282A
75 Watts	TW75, HF100, TF100, ZB120, HK257, 8001
80 Watts	828*
85 Watts	211 (Taylor), WE242A, GL242C
100 Watts	RK36, RK38, RK48, 100TH, 100TL, 203A, 211 (RCA), 211C, 211D, HK254, 261A, 276A, 813, GL835, 838, 845, 850, 852, 860, 8003
125 Watts	T125, 211C (Amperex), 211H (Amperex), 803, 805, GL146, GL152
150 Watts	TW150, HD203A, HK354, HK354C, D, E, F, 810*, 1627, 8000*
160 Watts	HF200
200 Watts	T200, HF300, T-814 (Taylor), T-822, HV12, HV18, HV27
225 Watts	806*
250 Watts	204A, 250TH, 250TL, GL159, GL169
275 Watts	212E
300 Watts	HK654
350 Watts	WE270A
400 Watts	831, 849, 861
450 Watts	450TH, 450TL, 833A*, HK854H, HK854L

**Intermittent-service rating. Continuous-service ratings run from 10 to 35 per cent lower than these figures.*

TRANSMITTING TUBE SOCKET CONNECTIONS—BOTTOM VIEWS.

REFERENCES INDICATED IN TUBE TABLES.

[1]Designed specifically for u.h.f. application.
[2]The suffix "H" indicates indirectly-heated cathode.
[3]S—small; M—medium; L—large. The final numbers refer to above socket connection diagrams.
[4]Grid driving requirements for r.f. service are subject to wide variation depending upon impedance of plate circuit. Values given are for typical plate impedances. A reserve of excitation power should always be available, and allowance should be made for appreciable circuit losses at ultra high frequencies when choosing a driver tube.
[5]Manufactured by the following: Amperex (Amp.), Eimac, General Electric Co. (G. E.), Heintz & Kaufman Ltd. (H & K), Hytron, Raytheon (Ray.), RCA Manufacturing Co. (RCA), Taylor, and United Electronics Co. (United).
[6]Intermittent commercial and amateur service ratings. For use where long tube life and reliability of operation are more important than tube cost, refer to more conservative ratings as given in manufacturer's data sheets.
[7]Plate current is the maximum signal value for Class B and Class AB audio applications.
[8]Grid current is the maximum signal value for Class B audio application.
[9]Plate-screen modulation is assumed in the Class C telephony application of tetrodes and pentodes.
[10]Bias must be adjusted at no signal for maximum rated dissipation.
[11]No-signal value for RK-100.
[12]Characteristics are per-section unless otherwise noted.
[13]Characteristics are for two tubes unless otherwise noted.
[14]Cathode connected to pin 4.
[15]Socket is provided with built-in by-pass condensers.
[16]Characteristics are for both sections unless otherwise noted.
[17]Grid connected to pins 2 and 3.
[18]Triode connected, screen tied to plate.
[19]At 112 Mc.

Mfr. No.	Type	Cathode Volts[2]	Cathode Amps.	Base[3] and Connections	Max. Plate Voltage	Max. Plate Current (ma.)	Triode Mu or Max. Screen Voltage	Max. D.C. Control Grid Current (ma.)	Max. Plate Dissipation (Watts)	Max. Screen Dissipation (Watts)	Interelectrode Capacitances $\mu\mu$fds. G-F Input	P-F Output	G-P Feedback	Typical[9] Operation	Plate Volts	Control Grid Bias	Screen Volts	Suppressor Volts	Plate[7] Current (ma.)	D.C.[8] Control Grid Current (ma.)	Screen Current (ma.)	[4]Grid Driving Power (Approx. Watts)	Load Imped. P to P (Ohms)	Power Output (Watts) typical	Mfr.[5]	Price
HY-6J5 GTX	Triode	6.3H	0.3	Cer.Octal 42	250	20	20	4.0	3.5		3.8	3.0	2.7	Class-C Telegraphy	250	−30			20	2.0		0.2		3	Hy-tron	.95
														Class-C Telephony	250	−30			20	2.5		0.4		3		
HY-6L6 GX	Beam Tetrode	6.3H	0.9	Cer.Octal 43	500	90	300	3.5	21	3.0	11.0	8.0	0.5	Class-C Telegraphy	500	−50	250		90	2.0	9	0.5		30	Hy-tron	1.25
														Class-C Telephony	400	−45	225		80	3.0	9	0.8		20		
														Class-AB$_2$ Audio	500	−25	300		230		20	0.6	4660	75		
HY-6V6 GTX	Beam Tetrode	6.3H	0.5	Cer. Octal 43	300	60	225	3.5	15	2.5	10.0	8.5	0.4	Class-C Telegraphy	300	−45	200		60	2.5	7.5	0.25		12	Hy-tron	1.05
														Class-C Telephony	250	−45	200		60	2.0	6	0.4		10		
														Class AB$_2$ Audio	300	−22.5	300		120		18	0.4	5000	22		
RK-10	Triode	7.5	1.25	4-pin. M. 3	450	65	8	15	15		3	4	8	Class-C Telegraphy	450	−100			65	15		3.2		19	Ray.	
														Class-C Telephony	350	−100			50	12		2.2		12		
RK-11	Triode	6.3	3.0	4-pin M. 6	750	105	20	35	25		7	0.9	7	Class-C Telegraphy	750	−120			105	21		3.2		55	Ray.	2.50
														Class-C Telephony	600	−120			85	24		3.7		38		
HV-12	Triode	10	4.0	Giant 4-Pin 29	2500	210	12	60	200		8.5	4	8.5	Class-C Telegraphy	2000	−300			200	9		8		300	United	18.00
														Class-B Audio	2000	−160			275			7	14400	400		
RK-12	Triode	6.3	3.0	4-pin M. 6	750	105	80	40	25		7	0.9	7	Class-C Telegraphy	750	−100			105	35		5.2		55	Ray.	2.50
														Class-C Telephony	600	−100			85	27		3.8		38		
														Class-B Audio	750	0			200	65		3.4	9600	100		
HK-14[1]	Triode	2.5	5.0	None (wire leads)	1500	50	25		15					U.H.F. Oscillator or Amplifier											H&K	
HV-18	Triode	10	3.85	Giant 4-Pin 13	2500	210	18	60	200		5	1.5	6.5	Class-C Telegraphy	2500	−300			200	18		8		375	United	22.50
														Class-C Telephony	2000	−350			160	20		9		250		
														Class-B Audio	2500	−130			360				16000	600		
RK-18	Triode	7.5	3.0	4-pin M. 6	1250	100	18	40	40		6	1.8	4.8	Class-C Telegraphy	1250	−160			100	12		2.8		95	Ray.	
														Class-C Telephony	1000	−140			80	13		3.1		64		
														Class-B Audio	1250	− 60			220	60		9	18000	190		

RK-19	Full-Wave Hi-Vacuum Rectifier	7.5H	2.5	4-pin M. 2	3500 Peak Inverse	600 Peak								Full-Wave Rectifier	1250				200						Ray.	
RK-20A	Pentode	7.5	3.25	5-pin M. 11	1250	92	300	15	40	15	14	12	0.01	Class-C Telegraphy	1250	−100	300	+45	92	11.5	36	1.6		84	Ray.	15.00
														Class-C Telephony	1000	−100	300	0	75	10	30	1.3		52		
T-20	Triode	7.5	1.75	4-pin M. 6	750	85	20	200	20		4.85	0.65	5.05	Class-C Telegraphy	750	− 85			85	18		3.6		44	Taylor	2.25
														Class-C Telephony	750	−135			70	15		3.6		38		
TZ-20	Triode	7.5	1.75	4-pin M. 6	750	85	62	30	20		5.25	0.55	4.95	Class-C Telegraphy	750	− 40			85	28		3.75		44	Taylor	2.25
														Class-C Telephony	750	−100			70	23		4.8		38		
														Class-B Audio	750	0			170			2.6	9000	80		
KY-21	Grid Control Mercury Vapor Rectifier	2.5	10	4-pin M. 6	11,000 Peak Inverse	3000 Peak								Grid-Controlled Rectifier					1500 (2 tubes)						Eimac	10.00
RK-21	Half-Wave Hi-Vacuum Rectifier	2.5H	4.0	4-pin M. 1	3500 Peak Inverse	600 Peak								DiodeHalf-Wave Hi-Vacuum Rectifier	1250 per plate				200						Ray.	
RX-21	Mercury Vapor Rectifier	2.5	10	4-pin M. 1	11,000 Peak Inverse	3000 Peak								Half-Wave Rectifier					1500 Output Current						Eimac	
T-21	Beam Power Tetrode	6.3H	0.9	6-pin M. 21	400	95	300	5	21	3.5	11.5	11.5	1.4	Class-C Telegraphy	400	−45	300		95	5	16	0.4		17	Taylor	1.95
														Class-C Telephony	400	−45	300		65	5	16	0.4		17		
RK-22	Full-Wave Hi-Vacuum Rectifier	2.5H	8.0	4-pin M. 2	3500 Peak Inverse	600 Peak								Full-Wave Rectifier	1250				200 Output Current						Ray.	
RK-23	Pentode	2.5H	2.0	7-pin M. 7	500	60	250	10	10	8.0	10	10	0.2	Class-C Telegraphy	500	−90	200	+45	55	4	38	0.5		22	Ray.	4.50
														Class-C Telephony	400	−90	150	0	43	6	30	0.8		13.5		
														Sup.-Mod. Telephony	500	−125	200	−45	31	4	39	0.5		6		
HK-24	Triode	6.3	3.0	4-pin S. 6	2000	75	25	25	25		2.5	0.4	1.7	Class-C Telegraphy	2000	−140			56	18		4.0		90	H&K	3.50
														Class-C Telephony	1500	−145			50	25		5.5		60		
														Class-B Audio	1250	−42			136			4.2		120		
HY-24	Triode	2.0	0.13	4-pin S. 3	180	20	9.3	4.5	2		2.7	2.3	5.4	Class-C Telegraphy	180	−45			20	4.5		0.2		2.7	Hy-tron	1.50
														Class-C Telephony	180	−45			20	4.5		0.3		2.5		
														Class-B Audio	180	−18			30			0.26	9200	2.7		

Mfr. No.	Type	Cathode Volts[2]	Cathode Amps.	Base[3] and Connections	Max. Plate Voltage	Max. Plate Current (ma.)	Triode Mu or Max. Screen Voltage	Max. D.C. Control Grid Current (ma.)	Max. Plate Dissipation (Watts)	Max. Screen Dissipation (Watts)	Interelectrode Capacitances μμfds. G-F Input	P-F Output	G-P Feedback	Typical[9] Operation	Plate Volts	Control Grid Bias	Screen Volts	Suppressor Volts	Plate[7] Current (ma.)	D.C.[8] Control Grid Current (ma.)	Screen Current (ma.)	[4]Grid Driving Power (Approx. Watts)	Load Imped. P to P (Ohms)	Power Output (Watts) typical	Mfr.[5]	Price
RK-24	Triode	2.0 d.c.	0.12	4-pin S 3	180	20.0	8	6.0	1.5		3.5	3.0	5.5	Class-C Amp.-Osc.	180	−45			16.5	6.0		0.5		2.0	Ray.	2.25
HY-25	Triode	7.5	2.25	4-pin M. 6	800	75	55	25	25		4.2	1.0	4.6	Class-C Telegraphy	750	−45			70	15		2.0		42	Hy-tron	1.45
														Class-C Telephony	700	−45			75	17		5.0		40		
														Class-B Audio	800	−9.0			140			2.7	2250	75		
RK-25 RK-25B	Pentode	6.3H	0.9	7-pin M. 7	500	60	250	10	10	8.0	10	10	0.2	Class-C Telegraphy	500	−90	200	+45	55	4	38	0.5		22	Ray.	3.95
														Class-C Telephony	400	−90	150	0	43	6	30	0.8		13.5		
														Sup.-Mod. Telephony	500	−125	200	−45	31	4	39	0.5		6		
HV-27	Triode	10	4	Giant 4-Pin 29	2500	210	26	60	200		8.5	3.5	14.5	Class–C Telegraphy	2000	−300			200	12		9		300	United	18.00
														Class–B Audio	2000	− 60			325			9	12400	400		
RK-28	Pentode	10	5.0	Giant 5-pin 11	2000	150	400	25	100	35	15	15	0.02	Class-C Telegraphy	2000	−100	400	+45	150	13	55	2.0		210	Ray.	28.50
														Class-C Telephony	1500	−100	400	+45	135	13	52	2.0		155		
														Sup.-Mod.Telephony	2000	−100	400	−45	85	13	65	1.8		60		
HY-30-Z	Triode	6.3	2.25	4-pin M. 6[17]	850	90	87	25	30		6.0	1.0	4.8	Class–C Telegraphy	850	−75			90	25		2.5		58	Hy-tron	2.75
														Class–C Telephony	700	−75			90	25		3.5		47		
														Class–B Audio	850	0			180			2	10000	110		
RK-30	Triode	7.5	3.25	4-pin M. 5	1250	80	15	25	35		2.75	2.75	2.5	Class-C Telegraphy	1250	−180			90	18		5.2		85	Ray.	
														Class-C Telephony	1000	−200			80	15		4.5		60		
														Class-B Audio	1250	−70			130	26		3.4	21000	106		
Twin 30[2]	Dual Triode	6	4	4-pin M.	1500	85 ea. Sect.	32	30 ea. Sect.	30 ea. Sect.		1.9	0.2	2	Class-C Telegraphy	1500	−100			150	40		15		175	Eimac	13.50
														Class C Telephony	1250	−100			135	40		15		125		
HY-31Z[16]	Twin Triode	6.3	2.5	4-pin M. 31	500	150	45	30	30		5 per	1.9 secti	5.5 on	Class-C Telegraphy	500	−45			150	25		2.5		56	Hy-tron	3.50
														Class–C Telephony	400	−45			150	30		3.5		45		
														Class–B Audio	500	0			150			1.8	7000	51		

RK-31	Triode	7.5	3.0	4-pin M. 6	1250	115	75	76	40		7	2.0	10	Class-C Telegraphy	1250	−80			100	30		3.9		90	Ray.	10.00
														Class-C Telephony	1000	−80			100	28		3.5		70		
														Class-B Audio	1250	0			220	76		4.4	18000	190		
RK-32	Triode	7.5	3.25	4-pin M. 5	1250	100	11	25	50		2.5	0.7	3.4	Class-C Telegraphy	1250	−225			100	14		4.8		90	Ray.	
														Class-C Telephony	1000	−310			100	21		8.7		70		
RK-33	Twin Triode	6.3	0.6	7-pin S 17	250	20 ea. Sect.	10	6.0	2.5	Left Triode	3.0	2.5	3.0	Class-C Amp.-Osc.	250	−60			20	6.0		0.54		3.5	Ray.	
										Right Triode	2.0	2.5	2.0													
RK-34[1]	Twin Triode	6.3	0.8	7-pin M. 22	300	80 both triodes	30	20	10 both triodes		4.2 ea.	0.8 ea.	2.7 ea.	P.P.Class-C Amp.-Osc.	300	−36			80	20		1.8		16	Ray.	3.50
														Class-B Audio	300	−15			70	12		0.5	10000	13		
35-T	Triode	5.0	4.0	4-pin M. 6	2000	150	30	35	70		4	0.2	1.9	Class-C Telegraphy	2000	−150			150	30		30		225	Eimac	6.00
														Class-C Telephony	1500	−120			100	30		15		120		
														Class-B Audio	1500	−40							12800	230		
35-TG	Triode	5.0	4.0	4-pin M. 5	2000	150	30	35	70		1.9	0.2	1.7	Same as 35-T											Eimac	6.75
RK-35	Triode	7.5	4.0	4-pin M. 5	1500	125	9	20	50		3.5	0.4	2.7	Class-C Telegraphy	1500	−250			115	15		5		120	Ray.	
														Class-C Telephony	1250	−250			100	14		4.6		93		
RK-36	Triode	5.0	8.0	4-pin M. 5	3000	165	14	35	100		4.5	1.0	5.0	Class-C Telegraphy	2000	−360			150	30		15		200	Ray.	
														Class-C Telephony	2000	−360			150	30		15		200		
RK-37	Triode	7.5	4.0	4-pin M. 5	1500	125	30	35	50		3.5	0.2	3.2	Class-C Telegraphy	1500	−130			115	30		7.0		122	Ray.	6.95
														Class-C Telephony	1250	−150			100	23		5.6		90		
														Class-B Audio	1250	−35			235	60		7.2	18000	200		
RK-38	Triode	5.0	8.0	4-pin M. 5	3000	165	30	40	100		4.6	0.9	4.3	Class-C Telegraphy	2000	−200			160	30		10		225	Ray.	13.50
														Class-C Telephony	2000	−200			160	30		10		225		
														Class-B Audio	2000	−52			265	39		5.8	16000	330		
RK-39	Beam Power Tetrode	6.3	0.9	5-pin M. 30	600	100	300	5.0	25	3.5	13	10	0.2	Class-C Telegraphy	600	−90	300		93	3.0	10	0.38		36	Ray.	3.50
														Class-C Telephony	475	−50	250		85	2.5	9.0	0.2		26		
														Class-AB_2 Audio	600	−30	300		200			0.4	6660	80		
HY-40	Triode	7.5	2.25	4-pin M. 6	1000	125	25	25	40		6.2	1.1	6.3	Class-C Telegraphy	1000	−90			125	20		5.0		94	Hy-tron	3.75
														Class-C Telephony	850	−90			125	15		3.5		82		
														Class-B Audio	1000	−22.5			250				9000	185		

Mfr. No.	Type	Cathode Volts[2]	Cathode Amps.	Base[3] and Connections	Max. Plate Voltage	Max. Plate Current (ma.)	Triode Mu or Max. Screen Voltage	Max. D.C. Control Grid Current (ma.)	Max. Plate Dissipation (Watts)	Max. Screen Dissipation (Watts)	Interelectrode Capacitances μμ fds. G-F Input	P-F Output	G P Feedback	Typical[9] Operation	Plate Volts	Control Grid Bias	Screen Volts	Suppressor Volts	Plate[7] Current (ma.)	D.C.[8] Control Grid Current (ma.)	Screen Current (ma.)	[4]Grid Driving Power (Approx. Watts)	Load Imped. P to P (Ohms)	Power Output (Watts) typical	Mfr.[5]	Price
HY-40Z	Triode	7.5	2.5	4-pin M. 6	1000	125	80	30	40		6.2	1.1	6.3	Class-C Telegraphy	1000	−27.5			125	25		5.0		94	Hy-tron	3.75
														Class-C Telephony	850	−30			125	30		7.0		82		
														Class-B Audio	1000	0			250	51		4.0	9000	185		
T-40	Triode	7.5	2.5	4-pin M. 6	1500	150	25	40	40		4.5	0.8	4.8	Class-C Telegraphy	1500	−140			150	28		9.0		158	Taylor	3.50
														Class-C Telephony	1250	−115			115	20		5.25		104		
TZ-40	Triode	7.5	2.5	4-pin M. 6	1500	150	62	45	40		4.8	0.8	5.0	Class-C Telegraphy	1500	−90			150	38		10		165	Taylor	3.50
														Class-C Telephony	1250	−100			125	30		7.5		116		
														Class-B Audio	1500	−9.0			250			6.0	12000	250		
RK-41	Beam Power Tetrode	2.5H	2.4	5-pin M. 30	600	100	300	5.0	25	3.5	13	10	0.2	Class-C Telegraphy	600	−90	300		93	3.0	10	0.38		36	Ray.	3.50
														Class-C Telephony	475	−50	250		85	2.5	9.0	0.2		26		
RK-42	Triode	1.5	0.06	4-pin S. 3	180	7.5	8				3	2.1	6.0	Class-A Audio	180	−13.5			3.9						Ray.	1.10
RK-43	Twin Triode	1.5	0.12	6-pin S.	135	15 (both triodes)	13	3.0			1.9	2.1	4.2	Class-C Amp.-Osc.	135	−20			14	3.0		0.2		1.25	Ray.	1.50
														Class-B Audio	135	−6.0			12.5	1.0		0.027	24000	0.95		
RK-44	Pentode	12.6H	0.7	7-pin M. 7	500	80	200	8.0	12	8.0	16	10	0.2	Class-C Telegraphy	500	−75	200	+40	60	4.0	15	0.4		22	Ray.	
														Class-C Telephony	400	−40	140	+40	45	5.0	20	0.3		11		
RK-45	Pentode	12.6H	0.45	7-pin M. 7	500	60	250	10	10	8.0	10	10	0.02	Class-C Telegraphy	500	−90	200	+45	55	4.0	38	0.5		22	Ray.	
														Class-C Telephony	400	−90	150	0	43	6.0	30	0.8		13.5		
RK-46	Pentode	12.6	2.5	5-pin M. 11	1250	92	300	15	40	15	14	12	0.1	Class-C Telegraphy	1250	−100	300	+45	92	11.6	36	1.6		84	Ray.	
														Class-C Telephony	1000	−100	300	0	75	10	30	1.3		52		
RK-47	Beam Power Tetrode	10	3.25	5-pin M. 10	1250	150	300	10	50	10	13	10	0.12	Class-C Telegraphy	1250	−70	300		138	7.0	14	1.0		120	Ray.	17.50
														Class-C Telephony	900	−120	250		90	7.5	23	1.2		55		
RK-48	Beam Power Tetrode	10	5.0	Giant 5-pin 10	2000	180	400	25	100	22	17	13	0.13	Class-C Telegraphy	2000	−100	400		180	6.5	40	1.0		250	Ray.	27.50
														Class-C Telephony	1500	−100	400		148	6.5	50	1.0		165		
RK-49	Beam Power Tetrode	6.3H	0.9	6-pin M. 21	400	100	300	6.0	21	3.5	11.5	10.6	1.4	Class-C Telegraphy	400	−50	250		95	3.0	8.0	0.2		25	Ray.	2.10
														Class-C Telephony	300	−45	200		60	5.0	15	0.3		13		

HY-51A	Triodes	7.5	3.5	4-pin M 6	1000	175	25	25	65		7.1	1.1	7.0	Class-C Telegraphy	1000	−75			175	20		7.5		131	Hy-tron	4.75
HY-51B		10.0	2.25											Class-C Telephony	800	−67.5			150	15		7.5		104		
														Class-B Audio	1000	−45			300			5.0	9000	285		
HY-51Z	Triode	7.5	3.5	4-pin M. 6[17]	1000	175	85	35	65		7.1	1.1	7.0	Class-C Telegraphy	1000	−22.5			175	35		10		131	Hy-tron	4.75
														Class-C Telephony	1000	−30			150	35		10		104		
														Class-B Audio	1250	0			300			4.0	2500	285		
RK-51	Triode	7.5	3.75	4-pin M. 6	1500	150	20	40	60		2.5		6.0	Class-C Telegraphy	1500	−250			150	31		10		170	Ray.	6.95
														Class-C Telephony	1250	−200			105	17		4.5		96		
RK-52	Triode	7.5	3.75	4-pin M. 6	1500	130	150	50	62.5		6.6	2.2	12	Class-C Telegraphy	1500	−120			130	40		7.0		135	Ray.	6.95
														Class-C Telephony	1250	−120			115	47		8.5		105		
														Class-B Audio	1250	0			300	100		7.5	10000	250		
HK-54	Triode	5.0	5.0	4-pin M. 5	3000	135	27	20	50		1.9	0.2	1.9	Class-C Telegraphy	2000	−269			130	20		9.0		210	H&K	6.75
														Class-C Telephony	2000	−269			130	20		9.0		210		
														Class-B Audio	1500	−45			198			8.0		200		
T-55	Triode	7.5	3.0	4-pin M. 6	1500	165	20	40			4.95	1.15	3.85	Class-C Telegraphy	1500	−140			165	20		5.6		183.5	Taylor	6.00
														Class-C Telephony	1500	−200			135	20		6.75		158		
RK-60	Full-Wave Hi-Vacuum Rectifier	5	3	4-pin M. 2	2120 Peak Inverse	250								Full-Wave Rectifier	750				250 average						Ray.	
HY-60	Beam Power Tetrode	6.3H	0.5	5-pin M. 30	425	60	225	5.0	15	2.5	10	8.5	0.1	Class-C Telegraphy	425	−45	225		60	2.5	7.0	0.25		16	Hy-tron	2.75
														Class-C Telephony	325	−45	225		60	2.0	8.5	0.2		10		
HY-61 807	Beam Power Tetrode	6.3H	0.9	5-pin M. 30	600	100	300	5.0	25	3.5	11	7.0	0.2	Class-C Telegraphy	600	−50	250		100	3.0	9.0	0.22		40	Hy-tron	3.50
														Class-C Telephony	475	−50	225		83	2.0	9.0	0.13		27.5		
														Class AB_2 P. P. Audio	600	−30	300		200		20	0.4	6660	80		
HY-63	Beam Tetrode	2.5 1.25	.1125 .225	Cer. Octal 44	250	25	180	2	3	0.6	9.5	7.4	0.15	Class-C Telegraphy	250	−22.5	135		25	2.0	4	0.2		4.3	Hy-tron	3.00
														Class-C Telephony	250	−35	135		20	2.0	3	0.2		3.5		
														Class-AB_2 Audio	250		180		50		6	0.2		7		
HY-65	Beam Tetrode	6.3	0.8	Cer. Octal 40	450	63	250	6	10	2.5	9.5	7.4	0.12	Class-C Telegraphy	450	−45	200		63	3	7	0.5		19	Hy-tron	3.00
														Class-C Telephony	350	−45	200		63	3	7	0.5		14		
HY-67	Beam Tetrode	6.3 12.6	4.0 2.0	5-pin M. 45	1250	175	300	15	65	10				Class-C Telegraphy	1250	−80	300		175	10	22.5	1.5		152	Hy-tron	7.75
														Class-C Telephony	1000	−150	300		145	14	17.5	2		101		

Mfr. No.	Type	Cathode Volts[2]	Cathode Amps.	Base[3] and Connections	Max. Plate Voltage	Max. Plate Current (ma.)	Triode Mu or Max. Screen Voltage	Max. D.C. Control Grid Current (ma.)	Max. Plate Dissipation (Watts)	Max. Screen Dissipation (Watts)	Interelectrode Capacitances μμfds. G-F Input	P-F Output	G-P Feedback	Typical[9] Operation	Plate Volts	Control Grid Bias	Screen Volts	Suppressor Volts	Plate[7] Current (ma.)	D.C.[8] Control Grid Current (ma.)	Screen Current (ma.)	[4]Grid Driving Power (Approx. Watts)	Load Imped. P to P (Ohms)	Power Output (Watts) typical	Mfr.[5]	Price
HY-69	Beam Power Tetrode	6.3	1.5	5-pin M. 10	600	100	300	7.5	40	5.0	15.3	7.3	0.19	Class-C Telegraphy	600	−60	250		100	4.0	12.5	0.25		42	Hy-tron	3.95
														Class-C Telephony	600	−60	250		100	4.0	10	0.25		42		
														Class AB_2 P. P. Audio	600	−35	300		240		29	0.7	4500	97		
75-T	Triode	5	6.5	4-pin M.	3000	250	10	30	75		2.2	0.3	2.3	Class-C Telegraphy	2000	−400			225	30				375	Eimac	9.00
														Class-C Telephony	2000	−400			200	30				300		
HY-75[1]	Triode	6.3	2.5	Octal 31	450	100	10	20	15		1.8	0.95		Class–C Telegraphy	450	− 60			82	15		1.5		21[19]	Hy-tron	3.95
														Class–C Telephony	450	− 60			62	20		2.5		16[19]		
TW-75	Triode	7.5	4.15	4-pin M.	2000	175	20	60	75		3.35	0.7	1.5	Class-C Telegraphy	2000	−175			150	37		12.7		225	Taylor	8.00
														Class-C Telephony	2000	−260			125	32		13.2		198		
100-TH	Triode	5.0	6.5	4-pin M. 5	3000	225	30	50	100		2.2	0.3	2.0	Class-C Telegraphy	3000	−210			125	50		30		300	Eimac	13.50
														Class-C Telephony	3000	−210			125	50		30		300		
														Class-B Audio[10]	3000								30000	500		
100-TL	Triode	5.0	6.5	4-pin M. 5	3000	225	12	35	100		2.0	4.0	2.3	Class-C Telegraphy	3000	−600			135	30		40		300	Eimac	13.50
														Class-C Telephony	3000	−600			135	30		40		300		
														Class-B Audio[10]	3000								30000	465		
HF-100 TF-100	Triode	10	2.0	4-pin M. 5	1750	150	23	30	75		3.5	1.4	4.5	Class-C Telegraphy	1500	−200			150	18		6.0		170	Amp. Tayl.	13.50 12.50
														Class-C Telephony	1250	−250			110	21		8.0		105		
														Class-B Audio	1750	−62			270			9.0	16000	350		
RK-100	Gas Triode	6.3H	0.9	6-pin M. 21	150	250	40	100.0	15		23	3.0	19	Class-C R.F. Amp.	110				185	40		2.1		12	Ray.	7.00
														Class-A Audio	110	−1.6			65[11]	8.5				4.2		
														Class-A P.P. Audio	110	−1.6			130[11]	17			2000	9		
HY-114[1]	Triode	1.4	0.12	Octal 31	180	15	20	3			1.2	0.6	1.7	Class–C Telephony	180				15	3				2	Hy-tron	
HY-114B	Triode	1.25	0.145	Octal 31	180	15	12	3	2		1.4	1.45	1.85	Class–C Telegraphy	180	−30			15	1.5		0.15		2[19]	Hy-tron	2.25
														Class–C Telephony	180	−30			15	1.5		0.25		2[19]		

ZB-120	Triode	10	2.0	4-pin Giant 12	1500	160	90	40	75		5.3	3.2	5.2	Class-C Telegraphy	1250	−135			160	23		5.5		145	Amp.	
														Class-C Telephony	1000	−150			120	21		5.0		95		
														Class-B Audio	1250	0			300			4.0	9000	245		
T-125	Triode	10	4.5	4-pin Giant 13	2500	250	25	70	125		6.3	1.3	6.0	Class-C Telegraphy	2500	−200			250	35		12.5		500	Taylor	13.50
														Class-C Telephony	2000	−165			250	35		12		375		
125-M	Beam Tetrode	5	6.5	5-pin Giant	3000	225		30	125		13	3.7	0.1											375	Eimac	28.50
GL-146	Triode	10	3.25	Special Large 4-pin 6	1500	175	78	60	125		7.2	3.9	9.2	Class–C Telegraphy	1250	−150			180	30				150	G.E.	18.00
														Class–C Telephony	1000	−200			160	40				100		
														Class–B Audio	1250	0			320				8400	250		
TW-150	Triode	10	4.1	4-pin 50-Watt 13	3000	200	35	60	150		3.9	0.8	2.0	Class-C Telegraphy	3000	−170			200	45		17		470	Taylor	15.00
														Class-C Telephony	3000	−260			165	40		17		400		
152-TL	Triode	5 or 10	13 or 6.5	Special	3000	500	10	75	150		5	0.75	5	Two parallel connected 75-T's in one envelope											Eimac	20.00
GL-152	Triode	10	3.25	Special Large 4-pin 6	1500	175	25	60	125		7	4	8.8	Class–C Telegraphy	1250	−150			180	30				150	G.E.	18.00
														Class–C Telephony	1000	−150			160	30				100		
														Class–B Audio	1250	−40			320				8400	250		
GL-159	Triode	10	9.6	Special Large 4-pin 6	2000	400	20	100	250		11	5	17.6	Class–C Telegraphy	2000	−200			400	17		6		620	G.E.	80.00
														Class–C Telephony	1500	−240			400	23		9		450		
														Class–B Audio	2000	−100			660				6880	900		
GL-169	Triode	10	9.6	Special Large 4-pin 6	2000	400	85	100	250		11.5	4.7	19	Class–C Telegraphy	2000	−100			400	42		10		620	G.E.	80.00
														Class–C Telephony	1500	−100			400	45		10		450		
														Class–B Audio	2000	−18			660			6	7000	900		
HF-200	Triode	10	3.4	4-pin Giant 13	2500	200	18	50	150		5.2	1.2	5.8	Class-C Telegraphy	2500	−300			200	18		8.0		380	Amp.	24.50
														Class-C Telephony	2000	−350			160	20		9.0		250		
														Class-B Audio	2500	−130			360			8.0	16000	600		
T-200	Triode	10	5.75	4-pin Giant 13	2500	350	17	60	200		9.5	1.6	7.9	Class-C Telegraphy	2500	−265			300	48		20		590	Taylor	21.50
														Class-C Telephony	2000	−220			250	41		15		390		
203-A	Triode	10	3.25	4-pin Giant 12	1250	150	25	60	100		6.5	5.5	14.5	Class-C Telegraphy	1250	−125			150	25		7.0		130	RCA G.E.	10.00
														Class-C Telephony	1000	−135			150	50		14		100		
														Class-B Audio	1250	−45			320			11	9000	260		

Mfr. No.	Type	Cathode Volts[2]	Cathode Amps.	Base[3] and Connections	Max. Plate Voltage	Max. Plate Current (ma.)	Triode Mu or Max. Screen Voltage	Max. D.C. Control Grid Current (ma.)	Max. Plate Dissipation (Watts)	Max. Screen Dissipation (Watts)	Interelectrode Capacitances $\mu\mu$ fds. G-F Input	P-F Output	G-P Feedback	Typical[9] Operation	Plate Volts	Control Grid Bias	Screen Volts	Suppressor Volts	Plate[7] Current (ma.)	D.C.[8] Control Grid Current (ma.)	Screen Current (ma.)	[4]Grid Driving Power (Approx. Watts)	Load Imped. P to F (Ohms)	Power Output (Watts) typical	Mfr.[5]	Price
T-203A	Triode	10	3.25	4-pin Giant 12	1250	175	25	60	100		8.0	7.0	14	Class-C Telegraphy	1250	−125			150	25		7.0		130	Taylor	10.00
														Class-C Telephony	1000	−135			150	50		14		100		
														Class-B Audio	1250	−45			320			11	9000	260		
HD-203A	Triode	10	4.0	4-pin Giant 29	1750	250	25	60	150				12	Class-C R.F. Amp.	1750				250	60					Taylor	14.50
														Class-B Audio	1750	−67.5			365				10,000	400		
203-Z	Triode	10	3.25	4-pin Giant 29	1250	175	85		65					Class-B Audio	1250	−4.5			350			6.75	8000	300	Taylor	8.00
204-A	Triode	11	3.85	Special 16	3000	275	23	80	250		12.5	2.3	15	Class-C Telegraphy	2500	−200			250	30		15		450	Amp. RCA Taylor G.E.	95.00 85.00 60.00 85.00
														Class-C Telephony	2000	−250			250	35		20		350		
														Class-B Audio	2000	−60			500			20	8800	600		
205-D	Triode	4.5	1.6	4-pin M. 3	400	50	7.3	10	14		5.2	3.3	4.8	Class–C Telegraphy	400	−112			45			1.5		10	W.E. United	5.00
														Class–C Telephony	350	−144			35			1.7		7		
														Class–A Audio	400	−29			30				7600	1.3		
211	Triode	10	3.25	4-pin Giant 12	1250	175	12	50	100		6.0	5.5	14.5	Class-C Telegraphy	1250	−225			150	18		7.0		130	RCA G.E. W.E.	10.00
														Class-C Telephony	1000	−260			150	35		14		100		
														Class-B Audio	1250	−100			320			8.0		260		
T-211	Triode	10	3.25	4-pin Giant 12	1250	175	12	50	100		7.0	6.0	14	Class-C Telegraphy	1250	−200			150	30		11		125	Taylor	10.00
														Class-C Telephony	1000	−175			150	30		10		100		
														Class-B Audio	1250	−80			300			25	8000	200		
211-C	Triode	10	3.25	4-pin Giant 12	1250	210	12	50	125		5.5	3.5	9.0	Class-C Telegraphy	1250	−250			200	10		3.5		170		
														Class-C Telephony	1250	−300			166	8.0		3.5		148	Amp.	17.50
														Class-B Audio	1250	−90			400			4.5	6700	320	United	
T-211-C	Triode	10	3.25	4-pin Giant 12	1250	175	12.5	60	100		6.0	6.5	9.0	Class-C Telegraphy	1250				125	60				100	Taylor	12.50
														Class-C Telephony	1000	−160			150	50				100		
T-211-D 211-D	Triode	10	3.25	4-pin Giant 12	1250	175	12	50	100		7.0	6.0	14	Class-C Telegraphy	1250	−200			150	30		11		125	Taylor Amp.	10.00
														Class-C Telephony	1000	−175			150	30		10		100		
														Class-B Audio	1250	−80			300			25	8000	200		

211-H HD-211C	Triode	10	3.25	4-pin Giant 29	1500	210	12.5	50	125		5.5	1.9	7.2	Class-C Telegraphy	1500	−300			200	10		4.0		220	Amp. Taylor	17 50 14.50
														Class-C Telephony	1250	−300			166	8.0		3.5		148		
														Class-B Audio	1500	−110			400			5.0	8200	400		
212E	Triode	14	6.0	4-pin W. E. 18	3000	350	16	75	275		14.9	8.6	18.8	Class-C Telegraphy	2000	−250			300	60		25		400	W.E. United	75.00
														Class-C Telephony	1500	−200			300	60		21		300		
														Class-B Audio	1500	−75			600			50	5900	500		
217-A	Half-Wave Hi-Vacuum Rectifier	10	3.25	4-pin Giant 26	3500 Inverse Peak	600 Peak								Half-Wave Rectifier	3500 In-verse Peak				200 aver-age						RCA	20.00
217-C	Half-Wave Hi-Vacuum Rectifier	10	3.25	4-pin Giant 14	7500 Inverse Peak	600 Peak								Half-Wave Rectifier	7500 In-verse Peak				150 aver-age						RCA G.E.	20.00
Z-225	Half-Wave Mercury-Vapor Rectifier	2.5	5	4-pin M. 1	10000 Inverse Peak	1000 Peak								Half-Wave Rectifier					250 aver-age						United	1.65
GL-242-C	Triode	10	3.25	4-pin Giant 12	1250	150	12.5	50	85		6.1	4.7	13	Class–C Telegraphy	1250	−225			150	20		7		130	G.E.	15.00
														Class–C Telephony	1000	−260			150	35		14		100		
														Class–B Audio	1250	−95			300			8	7600	200		
WE-242A	Triode	10	3.25	4-pin Giant 12	1250	150	12.5	50	85		6.5	4.0	13	Class-C Telegraphy	1250	−200			150	30		11		125	W.E.	
														Class-C Telephony	1000	−175			150	30		10		100		
														Class-B Audio	1250	−80			300			25	8000	200		
T-249-B	Half-Wave Mercury Vapor Rectifier	2.5	7.5	4-pin M. 1	10000 Peak Inverse	1500 Peak								In Single-Phase Full-Wave Rectifier	3530 R.M. S. In-put				750 Out-put Mils.						Taylor	5.00
250-TH	Triode	5.0	10.5	4-pin Giant 13	3000	350	32	100	250		3.5	0.3	3.3	Class-C Telegraphy	3000	−210			330	75		99		750	Eimac	24.50
														Class-C Telephony	3000	−210			330	75		99		750		
														Class-B Audio[10]	1250								3280	540		
250-TL	Triode	5.0	10.5	4-pin Giant 13	3000	350	13	50	250		3.0	0.5	3.5	Class-C Telegraphy	3000	−600			330	45		99		750	Eimac	24.50
														Class-C Telephony	3000	−600			330	45		99		750		
														Class-B Audio[10]	1250								3280	540		
HK-254	Triode	5.0	7.5	4-pin Giant 13	4000	200	25	40	100		3.3	1.1	3.4	Class-C R.F. Amp.	3000	−251			167	40		19		400	H.&K.	12.50
														R.F. Doubler	2000	−600			125	40		33		150		
														Class-B Audio	3000	−100			245			14	30000	550		

Mfr. No.	Type	Cathode Volts[2]	Cathode Amps.	Base[3] and Connections	Max. Plate Voltage	Max. Plate Current (ma.)	Triode Mu or Max. Screen Voltage	Max. D.C. Control Grid Current (ma.)	Max. Plate Dissipation (Watts)	Max. Screen Dissipation (Watts)	Interelectrode Capacitances μμ fds. G-F Input	P-F Output	G-P Feedback	Typical[9] Operation	Plate Volts	Control Grid Bias	Screen Volts	Suppressor Volts	Plate[7] Current (ma.)	D.C.[8] Control Grid Current (ma.)	Screen Current (ma.)	[4]Grid Driving Power (Approx. Watts)	Load Imped. P to P (Ohms)	Power Output (Watts) typical	Mfr.[5]	Price
WE-254B	Tetrode	7.5	3.25	4-pin M. 9	750	75	150	25	25	4.0	11.2	5.4	0.85	Class-C R.F. Amp.	750	−125	150		75					37.5	W.E.	
HK-257	Beam Power Pentode	5.0	7.5	7-pin Giant Bayonet 25	3000	150	500	25	75		13.8	6.7	0.04	Class-C Telegraphy	2000	−200	500	+60	150	6.0	11	1.4		230	H.& K.	22.50
														Class-C Telephony	1800	−130	400	+60	135	8.0	11	1.7		178		
														Sup.Mod.Telephony	2000	−130	500	−300	55	3.0	27	0.4		35		
261-A 276-A	Triode	10	3.25	4-pin Giant 12	1250	210	12.5	50	125		5.5	3.5	9	Class–C Telegraphy	1250	−250			200	10		3.5		170	Amperex	
														Class–C Telephony	1250	−300			166	8		3.5		148		
														Class–B Audio	1250	−90			400			4.5	6700	320		
WE-270A	Triode	10	9.75	Special 16	3000	375	16	75	350		18	2.0	21	Class-C Telegraphy	3000	−375			350	70		37		700	W.E.	
														Class-C Telephony	2250	−300			300	70		32		450		
														Class-B Audio	2000	−105			750			75	6000	850		
GL-276-A	Triode	10	3	4-pin Giant 12	1250	125	12	50	100		6	4	9	Class–C Telegraphy	1250	−225			125	20		7		100	G.E.	16.00
														Class–C Telephony	1000	−260			125	35		14		85		
														Class–B Audio	1250	−95			250			7.5	9000	175		
WE-282A	Tetrode	10	3.0	4-pin Giant 9	1000	100	150	50	70	5.0	12.2	6.8	0.2	Class-C R.F. Amp.	1000	−150	150		100					67	W.E.	
HF-300	Triode	11	4.0	4-pin Giant 13	3000	275	23	60	200		6.0	1.4	6.5	Class-C Telegraphy	2000	−200			275	36		13		410	Amp.	35.00
														Class-C Telephony	2000	−300			250	36		17		385		
														Class-B Audio	2000	−72			480			14	9600	650		
WE-300A	Triode	5.0	1.2	4-pin M. 3	450	100	3.8		40		9.0	4.3	15	Class-A Audio	450	−97			80					14.6	W.E.	10.10
														Class-A P.P. Audio	450				140					25		
WE-304B[1]	Triode	7.5	3.25	4-pin M. 5	1250	100	11	25	50		2.0	0.7	2.5	Class-C R.F. Amp.	1250	−225			100					85	W.E.	12.50
														Class-B Audio	1250	−110			100			10	14000	140		
304-TL	Triode	5 or 10	26 or 13	Special	3000	1000		150	300		10	1.5	10	Four parallel connected 75T's in one envelope											Eimac	65.00
WE305A[1]	Tetrode	10	3.1	4-pin M.	1000	125	200	40	60	6.0	10.5	5.4	0.14	Class-C R.F. Amp.	1000	−270	200		125					85	W.E.	
WE-307A	Pentode	5.5	1.0	5-pin M. 10	500	60	250	7.0	15	6.0	15	12	0.55	Class-C Telegraphy	500	−35	250	0	60	1.4	13			20	W.E.	
														Sup.-Mod. Telephony	500	−35	200	−50	40	1.5	20			6.0		

310	Triode	7.5	1.25	4-pin M. 3	600	70	8	15	20		4	2.2	7	Class-C Telegraphy	600	−150			65	15		4		25	United	3.45
														Class-C Telephony	500	−190			55	15		4.5		18		
														Class-B Audio	600	−75			130				10000	45		
WE316A[1]	Triode	2.0	3.65	No base	450	80		12	30		1.2	0.8	1.6	Class-C Telegraphy	450	Ad-justed for Tube			80					7.5 at 500 Mc.	W.E.	10.50
HK-354	Triode	5.0	10	4-pin Giant 29	4000	300	14	50	150		9.0	0.4	4.0	Class-C Telegraphy	4000	−690			245	50		48		830	H.& K.	24.50
														Class-C Telephony	3000	−550			210	50		35		525		
														Class-B Audio	2500	−165			350			20	15000	577		
HK354C[1]	Triode	5.0	10	4-pin Giant 13	4000	300	14	50	150		4.5	1.1	3.8	Class-C Telegraphy	4000	−690			245	50		48		830	H.& K.	24.50
														Class-C Telephony	3000	−550			210	50		35		525		
														Class-B Audio	2500	−165			350			20	15000	577		
HK-354D	Triode	5.0	10	4-pin Giant 13	3500	300	22	55	150		4.5	1.1	3.8	Class-C R.F. Amp.	3500	−490			240	50		38		690	H.& K.	24.50
														Class-B Audio	2500	−112			290			20	20000	519		
HK-354E	Triode	5.0	10	4-pin Giant 13	3500	300	35	60	150		4.5	1.1	3.8	Class-C R.F. Amp.	3500	−448			240	60		45		690	H.& K.	24.50
														Class-B Audio	2000	−37.5			372			20	11000	472		
HK-354F	Triode	5.0	10	4-pin Giant 13	3500	390	50	75	150		4.5	1.1	3.8	Class-C R.F. Amp.	3500	−368			250	75		50		720	H.& K.	24.50
														Class-B Audio	2500	−35			300			20	20000	550		
450-TH	Triode	7.5	12	4-pin Giant 13	6000	500	32	125	450		4.0	0.6	4.0	Class-C Telegraphy	4000	−400			500	70		100		1550	Eimac	75.00
														Class-C Telephony	4000	−400			400	70		100		1250		
450-TL	Triode	7.5	12	4-pin Giant 13	6000	500	16	75	450		4.0	0.6	4.0	Class-C Telegraphy	4000	−700			500	70		100		1550	Eimac	75.00
														Class-C Telephony	4000	−700			400	70		100		1250		
HY-615[1]	Triode	6.3H	0.15	Octal 32	300	20	22	4	3.5		1.4	1.45	1.85	Class-C Telegraphy	300	−35			20	1.4		0.2		4.5[19]	Hy-tron	2.25
														Class-C Telephony	300	−35			20	2.0		0.4		4.5[19]		
HK-654	Triode	7.5	15	4-pin Giant 13	4000	600	25	100	300		6.2	1.5	5.5	Class-C Telegraphy	2500	−406			500	75		59		950	H.& K.	75.00
														Class-C Telephony	3000	−390			400	95		60		945		
														Class-B Audio	1500	−45			643			50		675		
800	Triode	7.5	3.25	4-pin M. 5	1250	80	15		35		2.75	2.75	2.5	Class-C Telegraphy	1250	−175			70	15		4.0		65	RCA G.E.	10.00
														Class-C Telephony	1000	−200			70	15		4.0		50		
														Class-B Audio	1250	−70			130			3.4		106		

Mfr. No.	Type	Cathode Volts[2]	Cathode Amps.	Base[3] and Connections	Max. Plate Voltage	Max. Plate Current (ma.)	Triode Mu or Max. Screen Voltage	Max. D.C. Control Grid Current (ma.)	Max. Plate Dissipation (Watts)	Max Screen Dissipation (Watts)	Interelectrode Capacitances $\mu\mu$fds. G-F Input	P-F Output	G-P Feedback	Typical[9] Operation	Plate Volts	Control Grid Bias	Screen Volts	Suppressor Volts	Plate[7] Current (ma.)	D.C.[8] Control Grid Current (ma.)	Screen Current (ma.)	[4]Grid Driving Power (Approx. Watts)	Load Imped. P to P (Ohms)	Power Output (Watts) typical	Mfr.[5]	Price
801 801-A	Triode	7.5	1.25	4-pin M. 3	600	70	8		15		4.5	1.5	6.0	Class-C Telegraphy	600	−150			65	15		4.0		25	RCA G.E.	3.45
														Class-C Telephony	500	−190			55	15		4.5		18		
														Class-B Audio	600	−75			130			3.0		45		
HY 801-A 801	Triode	7.5	1.25	4-pin M. 3	600	70	8	15	20		4.5	1.5	6.0	Class-C Telegraphy	600	−200			70	15		4		30	Hy-tron	2.50
														Class-C Telephony	500	−200			60	15		4.5		22		
														Class-B Audio	600	−75			130	8		3	10000	22.5		
802[6]	Pentode	6.3H	0.9	7-pin M. 7	600	60	250	7.5	13		12	8.5	0.15	Class-C Telegraphy	500	−100	200	0	45	6.0	22	0.9		14	RCA G.E.	3.50
														Class-C Telephony	400	−40	200	+45	35	1.5	17	0.1		8		
														Sup.-Mod. Telephony	500	−45	200	−45	22	4.5	28	0.5		3.5		
803	Pentode	10	5.0	5-pin Giant 11	2000	175	600	50	125		17.5	29	0.15	Class-C Telegraphy	2000	−90	500	+40	160	12	45	2.0		210	RCA G.E.	28.50
														Class-C Telephony	1600	−80	500	+100	150	20	55	4.0		155		
														Sup.-Mod. Telephony	1500	−100	310	−90	100	20	70	3.5		50		
804[6]	Pentode	7.5	3.0	5-pin M. 11	1500	100	300	15	50		16	14.5	0.01	Class-C Telegraphy	1250	−100	300	+45	92	7.0	27	0.95		80	RCA	15.00
														Class-C Telephony	1000	−90	220	+50	75	6.0	21	0.65		50		
														Sup.-Mod. Telephony	1250	−100	300	−50	48	7.0	35.5	0.85		21		
805	Triode	10	3.25	4-pin Giant 29	1500	210	45	60	125		8.5	10.5	6.5	Class-C Telegraphy	1500	−105			200	40		8.5		215	RCA G.E.	13.50
														Class-C Telephony	1250	−160			160	60		16		140		
														Class-B Audio	1500	−16			400			7.0	8200	370		
T-805	Triode	10	3.2	4-pin Giant 29	1750	210	45	70	125		8.4	1.3	7.7	Class-C Telegraphy	1750	−90			200	44		9.2		270	Taylor	13.50
														Class-C Telephony	1500	−140			175	40		10.5		208		
														Class-B Audio	1750	−22.5			420			10	9350	320		
806[6]	Triode	5.0	10	4-pin Giant 29	3300	300	12.6	50	225		6.1	1.1	3.4	Class-C Telegraphy	3000	−600			195	25		20		450	RCA G.E.	22.00
														Class-C Telephony	2500	−600			195	40		32		390		
														Class-B Audio	2000	−150			390			14		500		
807[6]	Beam Power Tetrode	6.3H	0.9	5-pin M. 30	750	100	300	5.0	30		11	7.0	0.2	Class-C Telegraphy	600	−50	250		100	3.0	9.0	0.22		37.5	RCA G.E.	3.50
														Class-C Telephony	475	−50	225		83	2.0	9.0	0.13		24		
														Class AB_2 Audio	600	−30	300		200		10	0.4		80		

808	Triode	7.5	4.0	4-pin M. 5	1500	150	47	35	50		5.3	0.15	2.8	Class-C Telegraphy	1500	−200			125	30		9.5		140	RCA	7.75
														Class-C Telephony	1250	−225			100	32		10.5		105		
														Class-B Audio	1500	−25			190			4.8	18300	185		
809[6]	Triode	6.3	2.5	4-pin M. 6	1000	100	50	35	30		5.7	0.9	6.7	Class-C Telegraphy	750	−60			100	20		2.5		55	RCA G.E.	2.50
														Class-C Telephony	600	−160			83	32		7.2		38		
														Class-B Audio	750	−5			200			2.4	8400	100		
810[6]	Triode	10	4.5	4-pin Giant 29	2250	275	36	70	150		8.7	12	4.8	Class-C Telegraphy	2250	−160			275	40		12		475	RCA G.E.	13.50
														Class-C Telephony	1800	−200			250	50		17		335		
														Class-B Audio	2250	−60			450			13	11600	725		
811[6]	Triode	6.3	4.0	4-pin M. 6	1500	150	160	50	55		5.5	0.6	5.5	Class-C Telegraphy	1500	−113			150	35		8.0		170	RCA G.E.	3.50
														Class-C Telephony	1250	−125			125	50		11		120		
														Class-B Audio	1500	−9.0			200			4.2	18000	225		
812[6]	Triode	6.3	4.0	4-pin M. 6	1500	150	29	35	55		5.3	0.8	5.3	Class-C Telegraphy	1500	−175			150	25		6.5		170	RCA G.E.	3.50
														Class-C Telephony	1250	−125			125	25		6.0		120		
														Class-B Audio	1500	−46			200			4.7		225		
813	Beam Power Tetrode	10	5.0	7-pin L. 28	2000	180	400	25	100		16.3	14	0.2	Class-C Telegraphy	2000	−90	400		180	3.0	15	0.5		260	RCA G.E.	22.00
														Class-C Telephony	1600	−130	400		150	6.0	20	1.2		175		
814[6]	Beam Power Tetrode	10	3.25	5-pin M. 10	1500	150	300	15	65		13.5	13.5	0.1	Class-C Telegraphy	1500	−90	300		150	10	24	1.5		160	RCA G.E.	17.50
														Class-C Telephony	1250	−150	300		144	10	20	3.2		130		
T-814	Triode	10	4.0	4-pin Giant 29	3000	300	12	60	200		8.5	2.1	13.5	Class-C Telegraphy	2500	−190			300	51		17		600	Taylor	18.50
														Class-C Telephony	2000	−75			250	43		13.7		405		
														Class-B Audio	1500	−35			500			7.0	6800	525		
815[1,6]	P.P. Beam Power Tetrode	6.3 or 12.6	0.8 per Unit	Octal 27	500	150	200	10	25		13	9.5	0.15	Class-C Telegraphy	500	−100	200		150	10	17	0.13		56	RCA G.E.	
														Class-C Telephony	400	−100	200		150	10	15	0.16		45		
816	Half-Wave Mercury-Vapor Rectifier	2.5	2	4-pin M. 1	5000 Inverse Peak	500 Peak					3.2			Half-Wave Rectifier					125 Average						RCA	1.00
T-822	Triode	10	4.0	4-pin Giant 29	3000	300	30	60	200		8.5	2.1	13.5	Class-C Telegraphy	2500	−190			300	51		17		600	Taylor	18.50
														Class-C Telephony	2000	−75			250	43		13.7		405		
														Class-B Audio	1500	−35			500			7.0	6800	525		

Mfr. No.	Type	Cathode Volts[2]	Cathode Amps.	Base[3] and Connections	Max. Plate Voltage	Max. Plate Current (ma.)	Triode Mu or Max. Screen Voltage	Max. D.C. Control Grid Current (ma.)	Max. Plate Dissipation (Watts)	Max. Screen Dissipation (Watts)	Interelectrode Capacitances μμfds. G-F Input	P-F Output	G-P Feedback	Typical[9] Operation	Plate Volts	Control Grid Bias	Screen Volts	Suppressor Volts	Plate[7] Current (ma.)	D.C.[8] Control Grid Current (ma.)	Screen Current (ma.)	[4]Grid Driving Power (Approx. Watts)	Load Imped. P to P (Ohms)	Power Output (Watts) typical	Mfr.[5]	Price
825[1]	Inductive-Output Amplifier	6.3	0.75	Special Octal 33	2000 (Collector)	50 (Collector)		2.5	50 (Collector)					Special Amplifier, Multiplier, or Doubler for Frequencies Above 300 Mc.											RCA	34.50
826[1]	Triode	7.5	4	Special 34	1000	125	31	35	60		3.7	1.4	2.9	Class-C Telegraphy	1000	−70			125	35		5.8		86	RCA	19.00
														Class-C Telephony	800	−98			94	35		6.2		53		
827-R[1]	Beam Tetrode	7.5	25	Special 35	3500	500	1000	150	800		21	13	0.18	Class-C Telegraphy	3500	−300	700		428	100	185	50		1050	RCA	135.00
														Class-C Telephony	3000	−325	750		400	125	125	68		825		
828[6]	Beam Power Pentode	10	3.25	5-pin M. 10	2000	180	750	15	80		13.5	14.5	0.05	Class-C Telegraphy	1500	−100	400	+75	180	12	28	2.2		200	RCA	17.50
														Class-C Telephony	1250	−300	400	+75	160	12	28	2.7		150		
														Class AB_1 Audio	2000	−120	750	+60	270		60	0		385		
829[1]	P.P. Beam Power Tetrode[12]	6.3 or 12.6	1.125 per Unit	Special 24[15]	500	240	225	15	40		15.2	6.5	0.1	Class-C Telegraphy	500	−45	200		240	12	32	0.7		83	RCA G.E.	19.50
														Class-C Telephony	425	−60	200		212	11	35	0.8		63		
830-B	Triode	10	2.0	4-pin M. 6	1000	150	25	30	60		5.0	1.8	11	Class-C Telegraphy	1000	−110			140	30		7.0		90	RCA	10.00
														Class-C Telephony	800	−150			95	20		5.0		50		
														Class-B Audio	1000	−35			280			6.0	7600	175		
831[1]	Triode	11	10	Special Baseless	3500	350	14.5	75	400		3.8	1.4	4	Class-C Telegraphy	3500	−400			275	40		30		590	RCA	
														Class-C Telephony	3000	−500			200	60		50		360		
832[1]	P.P. Beam Power Tetrode[12]	6.3 or 12.6	0.8 per Unit	Special 24[15]	400	90		6.0	15		7.5	3.8	0.05	Class-C Telegraphy	400	−60	250		90	3.0	18	0.18		22	RCA	17.00
														Class-C Telephony	325	−50	210		68	1.2	15	0.06		12		
833-A[6]	Triode	10	10	Special 36	4000	500	35	100	450		12.3	8.5	6.3	Class-C Telegraphy	4000	−225			500	95		35		1600	RCA G.E.	85.00
														Class-C Telephony	4000	−325			450	90		42		1500		
														Class-B Audio	4000	−100			900			38	11000	2700		
834[1]	Triode	7.5	3.25	4-pin M. 5	1250	100	10.5	20	50		2.2	0.6	2.6	Class-C Telegraphy	1250	−225			90	15		4.5		75	RCA G.E.	12.50
														Class-C Telephony	1000	−310			90	17.5		6.5		58		
GL-835	Triode	10	3.25	Giant 4-pin 12	1250	175	12	50	100		6	5	9.25	Class-C Telegraphy	1250	−225			150	18		7		130	G.E.	16.00
														Class-C Telephony	1000	−260			150	35		14		100		
														Class-B Audio	1250	−95			320			8	9000	260		

836	Half-Wave Hi-Vacuum Rectifier	2.5H	5.0	4-pin M. 1	5000 Inverse Peak	1000 Peak								Half-Wave Rectifier					250 average						RCA G.E.	11.50
837	Pentode	12.6H	0.7	7-pin M. 7	500	50	200	8	12		16	10	0.2	Class-C Telegraphy	500	−75	200	+40	60	4.0	15	0.4		20	G.E. RCA	7.50
														Class-C Telephony	400	−40	140	+40	45	5.0	20	0.3		11		
838	Triode	10	3.25	4-pin Giant 12	1250	175	High	70	100		6.5	5.0	8.0	Class-C Telegraphy	1250	−90			150	30		6.0		130	RCA Taylor G.E.	11.00
														Class-C Telephony	1000	−135			150	60		16		100		
														Class-B Audio	1250	0			320			7.5	9000	260		
841	Triode	7.5	1.25	4-pin M. 3	450	60	30	20			4.0	3.0	7.0	Class-C Telegraphy	450	−34			50	15		1.8		15	RCA Ray.	3.25
														Class-C Telephony	350	−47			50	15		2.0		11		
														Class-B Audio	425	−5.0			114			3.6	7000	28		
843	Triode	2.5H	2.5	5-pin M. 4	450	40	7.7	7.5	15		4.0	4.0	4.5	Class-C Telegraphy	450	−140			30	5.0		1.0		7.5	G.E. RCA	12.50
														Class-C Telephony	350	−150			30	7.0		1.6		5.0		
844	Tetrode	2.5H	2.5	5-pin M. 8	500	30	180	5.0	15		9.5	7.5	0.15	Class-C Telegraphy	500	−125	175		25	5.0				9.0	RCA	18.00
														Class-C Telephony	500	−100	150		20					4.0		
845	Triode	10	3.25	4-pin Giant 12	1250	120	5.3		75		6.0	6.5	13.5	Class-A Audio	1250	−209			52					24	G.E. RCA Taylor	10.00
														Class-AB Audio	1250	−225			200				8800	105		
849	Triode	11	5.0	Special	3000	350	19	35	400		17	3.0	33.5	Class-C Telegraphy	2500	−250			300	20		8.0		560	G.E. RCA	120.00
														Class-C Telephony	2000	−300			300	30		14		425		
														Class-B Audio	2000	−105			650			16	6400	900		
849	Triode	11	5.0	Special	3000	350	19	125	400		11	2.0	33	Class-C Telegraphy	2500	−250			350	9.0		3.0		630	Amp.	160.00
														Class-C Telephony	2000	−300			300	10		4.0		425		
														Class-B Audio	2000	−90			680			8.0	6400	880		
852	Triode	10	3.25	4-pin M. top grid side plate	3000	150	12	40	100		1.9	1.0	2.6	Class-C Telegraphy	3000	−600			85	15		12		165	RCA	16.40
														Class-C Telephony	2000	−500			67	30		23		75		
HK-854H	Triode	7.5	12	Giant 4-pin 13	6000	600	30	110	450		8	0.5	4	Class-C Telegraphy	5000	−310			450	75		40		1820	H&K	
														Class-C Telephony	4000	−285			475	100		50		1520		
														Class-B Audio	4000	−140			670			45	14500	1970		
HK-854L	Triode	7.5	12	Giant 4-pin 13	6000	600	14	80	450		6	0.5	5	Class-C Telegraphy	5000	−575			450	45		40		1800	H&K	
														Class-C Telephony	4000	−625			475	65		58		1520		
														Class-B Audio	4000	−315			660			45	14500	1880		

Mfr. No.	Type	Cathode Volts[2]	Cathode Amps.	Base[3] and Connections	Max. Plate Voltage	Max. Plate Current (ma.)	Triode Mu or Max. Screen Voltage	Max. D.C. Control Grid Current (ma.)	Max. Plate Dissipation (Watts)	Max. Screen Dissipation (Watts)	Interelectrode Capacitances μμfds. G-F Input	P-F Output	G-P Feedback	Typical[9] Operation	Plate Volts	Control Grid Bias	Screen Volts	Suppressor Volts	Plate[7] Current (ma.)	D.C.[8] Control Grid Current (ma.)	Screen Current (ma.)	[4]Grid Driving Power (Approx. Watts)	Load Imped. P to P (Ohms)	Power Output (Watts) typical	Mfr.[5]	Price
860	Tetrode	10	3.25	4-pin M. top grid side plate	3000	150	300	40	100		7.75	7.5	0.08	Class-C Telegraphy	3000	−150	300		85	15		7.0		165	RCA G.E.	32.50
														Class-C Telephony	2000	−200	220		85	38	25	17		105		
861	Tetrode	11	10	Special	3500	350	500	75	400		14.5	10.5	0.1	Class-C Telegraphy	3500	−250	500		300	40	40	30		700	RCA G.E.	195.00
														Class-C Telephony	3000	−200	375		200	55		35		400		
865	Tetrode	7.5	2.0	4-pin M. 9	750	60		15	15		8.5	8	0.1	Class-C Telegraphy	750	−80	125		40	5.5		1.0		16	RCA G.E.	12.75
														Class-C Telephony	500	−120	125		40	9.0		2.5		10		
866 / 866-A	Half-Wave Mercury Vapor Rectifier	2.5	5.0	4-pin M. 1	10000 Inverse Peak	1000 Peak								Half-Wave Rectifier					250 average						Amp. Ray. RCA G.E. Hytron United	Av. 1.50
866 Jr.	Half-Wave Mercury Vapor Rectifier	2.5	3.0	4-pin M. 23	3500 Inverse Peak	250 Peak								Half-Wave Rectifier					125 average						Taylor	1.00
HY-866 Jr.	Half-Wave Mercury-Vapor Rectifier	2.5H	3.0	4-pin M. 23[14]	5000 Inverse Peak	500 Peak								Half-Wave Rectifier					125 Average						Hytron	1.05
872	Half-Wave Mercury Vapor Rectifier	5.0	10	4-pin Giant 14	7500 Inverse Peak	5000 Peak								Half-Wave Rectifier					1250 av.						RCA G.E.	9.00
872-A	Half-Wave Mercury Vapor Rectifier	5.0	6.75	4-pin Giant 14	10000 Inverse Peak	5000 Peak								Half-Wave Rectifier					1250 av.						Amp. Ray. RCA Taylor G.E.	Av. 11.00
HY-1231Z[16]	Twin Triode	12.6	1.5	5-pin M. 46	500	150	45	30	30		5.0 Per	1.9 Sect	5.5 ion	Class-C Telegraphy	500	−150			150	30		2.5		56	Hytron	4.50
														Class-C Telephony	400	−100			150	30		3.5		45		
														Class-B Audio	500	0			150	30		1.8	7000	51		
HY-1269	Beam Tetrode	12.6	1.5	5-pin M. 10	750	120	300	7.5	40	5	15.3	7.3	0.19	Class-C Telegraphy	750	−70	300		120	4	12.5	0.25		63	Hytron	4.50
														Class-C Telephony	600	−70	250		100	4	10.0	0.35		42		
														Class-AB_2 Audio	600	−35	300		240	6	29	0.7	4500	97		

975-A	Half-Wave Mercury-Vapor Rectifier	5	10	Giant 4-pin 14	15000 Inverse Peak	6000 Peak								Half-Wave Rectifier					1500 Average						United	21.00
1608	Triode	2.5	2.5	4-pin M. 3	425	95	20	25	20		8.5	3.0	9.0	Class-C Telegraphy	425	−200			95	25				20	RCA	4.00
														Class-C Telephony	350	−200			85	25				16		
1610	Crystal-Oscillator Pentode	2.5	1.75	5-pin M. 19	400	30	200	3	6	2	8.6	13	1.2	Oscillator-Amplifier	400	−50	150		22.5	1.5	7	0.1		5	RCA	2.00
1613	Pentode (Metal)	6.3H	0.7	Octal 47	350	50	275	5	10	2.5	8.5	11.5	0.5	Class-C Telegraphy	350	−35	200		50	3.5	10	0.22		9	RCA G.E.	2.75
														Class-C Telephony	275	−35	200		42	2.8	10	0.16		6		
1614	Beam Tetrode (Metal)	6.3H	0.9	Octal 43	375	110	300	5	21	315				Class-C Telegraphy	375	−35	200		88	3.5	9	0.18		17	RCA G.E.	3.50
														Class-C Telephony[18]	325	−70			65	9		0.8		11		
1619	Beam Tetrode (Metal)	2.5	2	Octal 37	400	75	300	5	15	3.5	10.5	12.5	0.35	Class-C Telegraphy	400	−55	300		75	5	10.5	0.36		19.5	RCA	4.75
														Class-C Telephony	325	−50	285		62	2.8	7.5	0.18		13		
														Class-AB_2 Audio	400	−16.5	300		150		11.5		6000	36		
1623	Triode	6.3	2.5	4-pin M. 6	1000	100	20	25	30		5.7	0.9	6.7	Class-C Telegraphy	750	−85			100	17		2.5		55	RCA G.E.	2.50
														Class-C Telephony	600	−125			83	25		5		38		
														Class-B Audio	750	−25			200			4.0	8400	100		
1624	Beam Power Tetrode	2.5	2.0	5-pin M. 10	600	90	300	5.0	25		11	7.5	0.25	Class-C Telegraphy	600	−60	300		90	5.0	10	0.43		35	RCA	3.50
														Class-C Telephony	500	−50	275		75	3.3	9.0	0.25		24		
														Class-AB_2 Audio	600	−25	300		42		15	1.2	7500	72		
1625	Beam Tetrode	12.6	0.45	7-pin M. 38	For Ot	her Ch	aracter	istics,	See 807	Data															RCA	3.50
1626	Triode	12.6	0.25	Octal 39	250	25	5	8	5		3.2	3.4	4.4	Oscillator	250	14000 ohms			25	5				4	RCA	2.50
1627	Triode	5	9	Giant 4-pin 29	For Ot	her Ch	aracter	istics,	See 810	Data															RCA	13.50
1628[1]	Triode	3.5	3.25	Special Base-less	1000	60	23	15	40		2.0	0.4	2.0	Class-C Telegraphy	1000	−65			50	15		1.7		35	RCA	32.00
														Class-C Telephony	800	−100			40	11		1.6		22		
8000[6]	Triode	10	4.5	Giant 4-pin 29	2250	275	16.5	40	150		5	3.3	6.4	Class-C Telegraphy	2250	−210			275	25		9		475	RCA	13.50
														Class-C Telephony	1800	−320			250	20		8.8		335		
														Self-Rectifying Osc.	2500 a.c.	5000 ohms	(Two	Tubes)	320	40				650		
														Class-B Audio	2250	−130			450			7.9	12000	725		

Mfr. No.	Type	Cathode Volts[2]	Cathode Amps.	Base[3] and Connections	Max. Plate Voltage	Max. Plate Current (ma.)	Triode Mu or Max. Screen Voltage	Max. D.C. Control Grid Current (ma.)	Max. Plate Dissipation (Watts)	Max. Screen Dissipation (Watts)	Interelectrode Capacitances μμfds. G-F Input	P-F Output	G-P Feedback	Typical[9] Operation	Plate Volts	Control Grid Bias	Screen Volts	Suppressor Volts	Plate[7] Current (ma.)	D.C.[8] Control Grid Current (ma.)	Screen Current (ma.)	[4]Grid Driving Power (Approx. Watts)	Load Imped. P to P (Ohms)	Power Output (Watts) typical	Mfr.[5]	Price
8001	Beam Pentode	5	7.5	Giant 7-pin 25	2000	150	500	25	75		11	5.5	0.1	Class-C Telegraphy	2000	−200	500	+60	150	6	11	1.4		230	RCA	22.00
														Class-C Telephony	1800	−130	400	+60	135	8	11	1.7		178		
														Sup. Mod. Telephony	2000	−130	500	−300	55	3	27	0.4		35		
8003	Triode	10	3.25	Giant 4-pin 29	1350	250	12	50	100		5.8	3.4	11.7	Class-C Telegraphy	1350	−180			245	35		11		250	RCA	12.00
														Class-C Telephony	1100	−260			200	40		15		167		
														Self-Rectifying Osc.	1500 a.c.	5000 ohms	(Two	Tubes)	400	40				500		
														Class-B Audio	1350	−100			490			10.5	6000	460		

Transmitter Design

RECEIVERS are designed pretty much as an integral unit, but there are infinite combinations of tubes, exciter circuits, amplifier circuits, and power supply arrangements which one may incorporate in a "200-watt" transmitter. For this reason, few complete transmitter circuit diagrams are shown in this book.

If a tube requires 25 watts r.f. driving power for a certain application, it is obvious that it makes little difference just what exciter circuit is used so long as it puts out 25 watts on the desired bands. Because of its characteristics one exciter may be preferred by one amateur, another exciter by another amateur.

It is fortunate that there is this flexibility with regard to transmitter design, because it makes it easy for an amateur to start out with a low power transmitter and then add to it from time to time, perhaps later going on phone. It also permits one a certain degree of "custom tailoring" of his transmitter to suit his particular requirements.

In several following chapters of this book are described inexpensive yet versatile and efficient exciters, power amplifiers, speech amplifiers, modulators, and power supplies. It is the purpose of this chapter to give the reader sufficient general design information to be able to work out various combinations of these independent yet complementary units, and to evolve one which is well suited to his particular needs and pocketbook. However, before proceeding further, one should be thoroughly familiar with the chapter on fundamental transmitter theory, Chapter 7.

Exciters and Transmitters. A 5-watt crystal oscillator may be accurately referred to as a transmitter *when it is used to feed an antenna.* On the other hand a multi-tube r.f. unit winding up in a 150-watt power amplifier may be properly termed an exciter *when it is used to drive a higher power amplifier.* Thus we see that any r.f. unit, even a simple oscillator, may be either an exciter or a transmitter depending upon how it is used.

The requirements for a low power (15 to 75 watt) transmitter are practically the same as for an exciter of the same output: The overall efficiency should be good, the unit should cover all the desired bands with a minimum of coil changing and retuning, and both initial cost and upkeep should be low in proportion to the power output.

Virtually all medium- and high-power amplifiers (200–800 watts output) are very much the same except for the particular make and power rating of components used. Perhaps half the amateurs making use of high power use cross-neutralized push-pull final amplifiers which differ only in the method of obtaining bias and method of antenna coupling.

For this reason, several low power r.f. units and several medium- and high-power amplifiers are described, and the reader is permitted to use his own ingenuity in working out the combination which appears to fit his requirements. If one is designing a complete transmitter, to which no additions are to be made, it is probably best to decide first upon the final amplifier and then work backwards from there, the driving requirements of the particular tubes used determining the exciter. On the other hand, many amateurs do not have the wherewithal to start right off with high power, and are therefore very likely to decide upon the highest powered r.f. unit they can afford and let it go at that. In the latter event, the unit may have slightly more output than is required to drive an amplifier whose addition is contemplated at a later date. However, a reserve of excitation power is not a liability and does not represent poor economy unless carried to extremes. Hence, one who cannot afford to start off with high power can pick out the highest powered exciter he can afford and use it as a transmitter, without worrying too much about its adaptability for use with a particular power amplifier later on. A 75-watt r.f. unit is slightly larger than necessary for driving a pair of 35T's, HK54's, 808's, T40's, HY51's, etc., but there is no reason why one should not use such a combination. *Not enough* excitation is a much more serious condition than an overabundance of excitation, there being no objec-

tion to the latter except from an economic standpoint.

Choosing the Tubes. Low-power exciters invariably use receiving tubes or "modified" receiving tubes for the sake of economy. Large scale production brings the cost of 42's, 6L6's, etc., down to a price that would be impossible were they designed for and purchased only by amateurs. Some tubes, like the T21 and 807, resemble standard receiving tubes in one or more respects, and while costing more than a standard receiving tube equivalent (6L6G in this case), are still obtainable at a price below that which would be necessary were they not outgrowths of receiving tubes.

The tubes in the high power amplifier and in the class B modulator (if used) should be chosen with care. While in general there is little to choose between tubes by reliable manufacturers, some are better adapted than others for certain applications. Also, the more recently released tubes of a particular manufacturer are usually better and less expensive than older tubes of the same general type.

Some of the older type tubes, such as the venerable 203-A, have been improved upon and their price periodically lowered until they compare favorably with recently released tubes; for certain applications they are a good buy and are highly recommended. Other tubes of this vintage, such as the 865, have been superseded by less expensive tubes giving better performance, and are manufactured primarily for replacement purposes.

Tubes for modulator service should have good emission and plate dissipation. Interelectrode capacities are relatively unimportant (within reason). For triode class B modulator service, the usual practice is to use high μ tubes so that little or no bias is required.

For oscillator service, tubes with medium μ are most satisfactory.

For doubler service, either pentodes, tetrodes, or high μ triodes are satisfactory.

For class C or class B r.f. service, the amplification factor is not important, though tubes with a medium high μ (20 to 30) are most popular.

In class A audio service, low μ triodes are to be preferred, though pentodes or beam tubes may be used when the load is constant or if inverse feedback is used.

Driving Power. It is always advisable to have a slight reserve of driving power in order to be on the safe side. Therefore, the potential output of an exciter on the band upon which its output is least (usually the highest frequency band) should be slightly greater than the excitation requirements of the following stage as determined from the manufacturer's tube data.

Plate modulated class C amplifiers require the most excitation, the tube requiring full maximum rated grid current and at least $2\frac{1}{2}$ times cutoff bias if full plate input is run.

C.w. and buffer amplifiers should preferably be run at full rated grid current (though they *may* run with as much as 50 per cent less) and at $1\frac{1}{2}$ times cutoff or greater bias. Thus an unmodulated final amplifier or buffer can be used with considerably less excitation than a plate modulated stage of the same power.

Cathode modulated amplifiers require about the same amount of excitation power as c.w. amplifiers, the bias being greater but the grid current much less. Cathode modulated stages are commonly run at from $2\frac{1}{2}$ to 4 times cutoff bias at approximately an eighth the grid current recommended for plate modulation.

High efficiency grid modulation requires still less excitation. The bias is from 2 to 4 times cutoff but the grid current is very low, seldom greater than a few ma. even for high power stages. The power dissipated in the grid swamping resistor, a necessary adjunct to a correctly operated grid modulated stage, keeps the excitation requirements from being less than they are.

The excitation required for a typical 200-watt output amplifier will run about as follows: plate modulated, 35 watts; c.w. or buffer, 20 watts; cathode modulated, 15 watts; grid modulated, 8 watts. The whole problem of excitation requirements depends so much upon operating conditions that one had best refer to the manufacturers' data sheets or to Chapter 10 of this Handbook.

The question of calculating excitation requirements for a doubler stage was not covered in the foregoing discussion, because the excitation power required depends to such a great degree upon the doubler efficiency desired. For high efficiency doublers, the bias should be at least 5 times cutoff and the grid current about half the maximum rated value for the tube. Thus it is seen that for good doubler efficiency a tube requires as much excitation power as does a plate modulated stage of the same power output rating.

Also to be taken into consideration, when tentatively planning a transmitter, are such things as the limiting factor in tube design. For instance, in a grid modulated transmitter, the ouput is always limited by the plate dissipation, while for plate modulated phone work either the plate voltage or plate current rating is exceeded first. Thus we see that for grid modulation, a tube with high plate dissipation, is of prime importance, while for plate modulated operation the matter of filament emission and insulation are of greatest importance.

Another thing to be taken into consideration.

especially when designing a phone transmitter, is the item of filament voltage. Obviously a saving can be effected if both r.f. amplifier tubes and modulator tubes can be run from the same filament winding.

Care should be taken to make sure that the tubes chosen are capable of efficient and safe operation on the highest frequency used.

Design Considerations

Transmitter Wiring. At the higher frequencies, solid enamelled copper wire is most efficient for r.f. leads. Tinned or stranded wire will show greater losses at these frequencies. Tank coil and tank condenser leads should be of heavier wire than other r.f. leads, though there is little point in using wire heavier than is used for the tank coil itself.

All grounds and by-passes in an r.f. stage should be made to a common point, and the grounding points for several stages bonded together with heavy wire.

Best type flexible leads to terminals coming out envelope of tubes is thin copper strip, cut from thin sheet copper. Heavy, rigid leads to these terminals may crack the envelope glass when a tube heats or cools.

Wires carrying only a.f. or d.c. should be chosen with the voltage and current in mind. Some of the low-voltage-filament type transmitting tubes draw heavy current, and heavy wire must be used to avoid voltage drop. The voltage is low, and, hence, not much insulation is required. Filament and heater leads are usually twisted together. An initial check should be made on the filament voltage of all tubes of 25 watts or more plate dissipation rating. This voltage should be measured right at the tube sockets. If it is low, the filament transformer voltage should be raised. If this is impossible, heavier or paralleled wires should be used for filament leads, cutting down their length if possible.

Spark plug type high tension ignition cable makes the best wire for high voltage leads. This cable will safely withstand the highest voltages encountered in an amateur transmitter. If this cable is used, the high voltage leads may be cabled right in with filament and other low voltage leads. For high voltage leads in low-power exciters, where the plate voltage is not over 450 volts, ordinary radio hookup wire of good quality will serve the purpose. Twisted lamp cord, in good condition with insulation intact, can be used for power supply leads between low-power exciter units and power supplies where the voltage does not exceed 400 volts.

No r.f. leads should be cabled; in fact it is better to use enamelled or bare copper wire for r.f. leads and rely upon spacing for insulation. All r.f. joints should be soldered, and the joint should be a good mechanical junction before solder is applied. Soldering technique is covered in Chapter 25.

Coil Placement. While metal shield baffles are effective in suppressing stray capacity coupling between circuits, they are not always effective in suppressing inductive coupling. To eliminate all inductive coupling between two coils in inductive relation to each other, each coil should be completely enclosed in an individual shield can. This is not always convenient; so more often the inductive coupling is minimized by orienting the coils for maximum suppression of coupling, and shield baffles are used only to prevent stray capacity coupling between stages.

For best Q a coil should be in the form of a solenoid approximately as long as its diameter. For minimum interstage coupling, coils should be made as small physically as is practicable. The coils should then be placed so that adjoining coils are oriented for minimum mutual coupling. To determine if this condition exists, apply the following test: the axis of one of the two coils must lie in the plane formed by the center turn of the other coil. If this condition is not met, there is bound to be appreciable coupling unless the unshielded coils are very small in diameter or are spaced a considerable distance from each other.

Variable Condensers. The question of optimum C/L ratio and condenser plate spacing is covered in the chapter on transmitter theory. For all-band operation of a high power stage, it is recommended that a condenser just large enough for 40-meter c.w. operation be chosen. (This will have sufficient capacity for 'phone operation on all higher frequency bands.) Then use fixed padding condensers for operation on 80 and 160 meters. Such padding condensers are available in air, gas-filled, and vacuum types.

Specially designed variable condensers are recommended for u.h.f. work; ordinary condensers often have "loops" in the metal frame which resonate near the operating frequency.

Insulation. On frequencies above 7 Mc., ceramic, polystyrene, or Mycalex insulation is to be recommended, though hard rubber will do almost as well. Cold flow must be considered when using polystyrene (Victron, Amphenol 912, etc.) or hard rubber. Bakelite has low losses on the lower frequencies but should never be used in the field of high-frequency tank circuits.

Lucite, which is available in rods, sheets, or tubing, is excellent for use at all radio frequencies where the r.f. voltages are not especially high. It is very easy to work with ordi-

nary tools and is not expensive. The loss factor depends to a considerable extent upon the amount and kind of plasticizer used.

The most important thing to keep in mind regarding insulation is that the best insulation is none at all. If it is necessary to reinforce air-wound coils to keep turns from vibrating or touching, use strips of Lucite or polystyrene cemented in place with Amphenol 912 coil dope. This will result in lower losses than the commonly used celluloid ribs and Duco cement.

Metering. The ideal transmitter would have an individual meter in every circuit requiring measurement. However, for the sake of economy, many of us are forced to measure filament and plate voltages by means of a test set or universal meter during the initial tryout of the transmitter, and then assume that these voltages will be maintained. Further economies can be effected by doubling up on meters when measuring current in various circuits in which the current is variable, and is an index of transmitter tuning.

By a system of plugs and jacks, or a selector switch, one or two milliammeters can be used to make all the measurements necessary to tune up a transmitter properly. However, it often is of considerable advantage to be able to observe the current of several circuits or stages simultaneously. Thus the problem boils down to: buy as many meters as you can afford, or as many as the total transmitter investment justifies, purchasing the most necessary meters first. Obviously one would not be justified in buying $100 worth of meters for a transmitter containing other parts totaling $75. On the other hand, the purchase of a filament voltmeter to keep careful tab on the filament voltage of a pair of 250 watt tubes is a good investment.

Probably the most popular arrangement calls for meter switching or meter jacks in the low power stages and individual meters in the last stage. Ordinarily, r.f. meters are not used except in certain antenna coupling circuits. Where line voltage does not fluctuate appreciably, one can get by very nicely with just d.c. milliammeters, plate current meters in the low power stages, and a grid and a plate meter in the final stage.

Where it is impossible to keep meter or meter leads well away from high r.f. voltage or heavy r.f. current, d.c. meters should be by-passed with small .004 or larger condensers directly at the meter terminals. The condenser is placed across the terminals, not from one terminal to ground. Such condensers are a wise precaution in all cases, because even though meter and meter leads are kept away from r.f. components, the meter may be subjected to considerable r.f. because of an r.f. choke not doing a 100 per cent job of blocking r.f. from the meter.

Most meters now come with bakelite cases. If the "zero adjuster" screw is well insulated, such meters can be placed in positive high voltage leads where the voltage does not exceed 1000 volts. When the voltage is higher than 1000 volts, the meter should preferably be placed behind a protective glass. The meter should not be mounted directly on a grounded metal panel when the plate voltage exceeds 2000 volts, as the metal portions of the meter may arc through the bakelite case to the grounded metal panel, particularly when plate modulation is used.

One highly recommended method of arranging meters in a high-powered rack and panel transmitter is to group all meters on a Masonite meter panel at the top of the rack, near eye level of the operator and not close to any of the tuning dials. With the Masonite meter panel, there is no danger of meters arcing to ground, and because of the position of the meters there is little likelihood of an operator accidentally coming in contact with the meters.

An alternate system is to place all meters in low voltage circuits directly on the metal panels (assuming meters are of the bakelite case type) and to place the plate milliammeters in all stages having a plate voltage of more than 1000 behind the panel, where they are observed through small windows.

Meter Switching. This method can be used to advantage where the voltages on the leads which carry the current to be measured are not greater than about 500 volts to ground. Fifty-ohm resistors are inserted in the leads, and because the resistance of the meter is so low compared to the 50-ohm resistors, the meter can be considered as being inserted in series with the circuit when it is tapped across the resistor. Thus, with a double pole selector switch having sufficient positions, one can use a single meter to measure the current in several circuits.

The resistor should be made 25 ohms where the current to be measured runs over 200 ma., and the resistor increased to 200 ohms when the current to be measured is less than 15 ma. It is necessary to minimize the resistance where heavy current is present, in order to avoid excessive voltage drop when the meter is not shunting the resistor. It is necessary to increase the value of resistance when the current is so low that a low range meter must be used to measure the current. Low range milliammeters begin to show appreciable resistance themselves, and their calibration will be thrown off when shunted by too low a value of resistor.

Meter switching is not practicable in high

voltage circuits (over 1200 volts). For measuring plate current in high power stages, the resistor should be placed either in the B minus lead or in the filament return (center tap). Placing the meter resistor in the B minus is not practical except when a power supply is used to feed but a single stage, or when heater-type tubes or separate filament transformers are used, as otherwise the meter would indicate total current to all the stages.

Placing the meter in the filament return gives a reading of the total *space current,* which includes both grid current and plate current (and in the case of tetrodes and pentodes, screen current). This point is covered later under *Meter Jacks.*

It is possible, by means of various systems of shunts, to use a single low-range meter for measuring widely different values of current in different circuits, much in the manner of the single-meter test set so popular with servicemen. For instance, a 0–25 ma. meter could be used for measuring grid current in several stages, and then used as a 0–250 ma. instrument when switched into the plate circuit of the final stage by the incorporation of a shunt in the latter circuit to extend the range to 250 ma. Ordinarily, however, a meter is used as a single-scale instrument with this type of switching, a 0–25 ma. meter being used only to read current in circuits carrying up to 25 ma.

Meter Jacks. A popular method of using one meter to measure the current in several circuits is to incorporate jacks in the various circuits to be measured. Instead of using low values of resistors across the packs to provide a current path when the meter is not plugged in a circuit, shorting type jacks are used so that when a meter is removed from a jack the circuit is automatically closed.

As with meter switching, meter shunts may be placed across certain of the jacks to extend the range of a milliammeter; however, it is more common practice to have a low range meter and a high range meter, and plug the appropriate meter in each circuit.

Meter jacks should not be used except where one side of the circuit can be grounded. This permits one to measure grid current, and, indirectly, plate current. The plate current is ascertained by measuring the current flowing in the filament return and subtracting the grid current (including screen current if the tube has a screen).

In connecting up meter jacks it is important that they be wired so that the meters read in the correct direction. This can be determined by figuring just which way the current is flowing in each circuit. If this were not done, the leads to the meter would have to be reversed when reading grid current after cathode current.

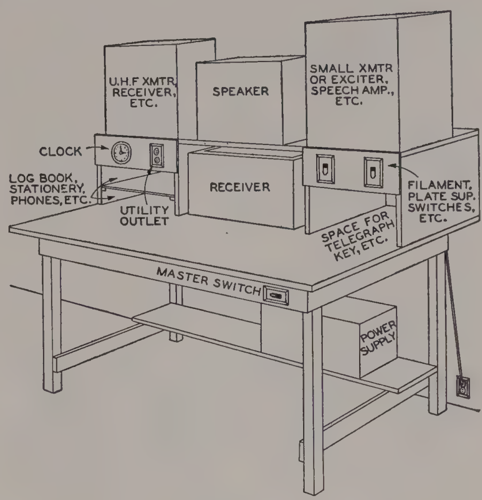


Figure 1.

UTILITY TYPE OPERATING TABLE.

Any amateur handy with a hammer and saw can construct a table of this type with little difficulty and at small cost. If a power supply is placed under the table as shown, it should be housed on the front and top in order to protect the operator from accidental contact with any of the components with his foot. If the equipment supported by the table is especially heavy, the two back legs of the table should be cross braced.

It necessitates insulating the frame of either the grid current jack or the cathode current jack from a grounded metal panel if such a panel is used. It is common practice to ground the frame of the cathode circuit jack and insulate the frame of the grid current jack, as this affords maximum protection to the operator.

A piece of heavily-insulated rubber covered 2-wire cable can be used to connect the meter to the meter plug. If the meter is permanently mounted on the panel, the meter cord should be long enough to reach all meter jacks into which it is to be plugged. To protect low range meters, cathode current jacks in stages drawing heavy current are usually placed in such a position that it is impossible to reach the jack with the cord attached to the low-range meter.

Meter jacks should never be placed in high-voltage leads, and it is inadvisable to use them in any circuit where one side of the jack is not at ground potential. When used for measuring cathode current, the *frame* of the jack should always be grounded, as a defective contact in the jack or a blown meter might

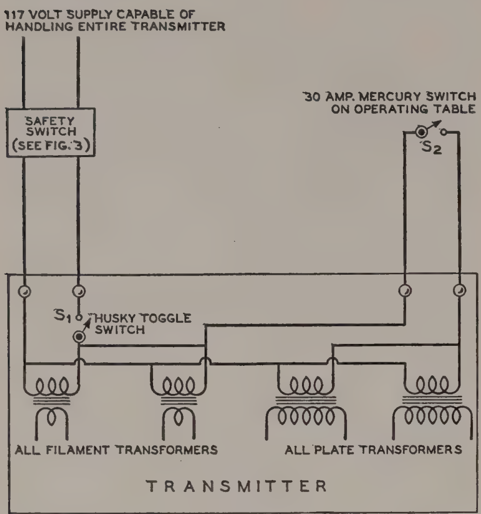


Figure 2.

POPULAR METHOD OF SUPPLYING AND SWITCHING MEDIUM POWER TRANSMITTER.

This arrangement is the one most widely used when the transmitter cannot be reached easily from the operating position. S_2 should never be turned on until S_1 has first been on for 15 seconds, preferably 30 seconds. S_1 should never be turned on except when S_2 is off; thus S_2 should always be turned off before S_1 is turned off.

otherwise endanger the operator by putting high potential on the meter cord and plug.

A 50-ohm carbon resistor across the terminals of all cathode current meter jacks will not affect the calibration of the meter, yet will protect the operator from possible shock in the event that the meter should blow or the cord open up or come loose on the ground side. In this case, the resistor is more of a protective device than a substitute path for the current when the meter is being used in some other circuit, and little current will flow through the resistor unless the jack, cord, or meter becomes defective.

The Audio System. In constructing audio equipment, the low level stages should always be mounted on a metal chassis and the bottom of the chassis shielded. For amateur work, "high fidelity" is neither necessary nor desirable, as the sideband width is increased without an increase in intelligibility. This means that high-output microphones of the "p.a." type designed particularly for speech transmission (such as the high output, diaphragm crystal) can be used, and the speech amplifier need have but moderate gain. This greatly simplifies the problem of construction, as the difficulties and chances for trouble go up rapidly as the maximum overall gain of an amplifier is increased much beyond this point. Elaborate precautions against r.f. and a.f. feedback and hum pickup must be taken when low-level high-fidelity microphones of the broadcast type are used, but with the type recommended only a few simple precautions need be taken.

If a microphone which requires an input transformer is used, such as the dynamic type, care must be taken in the orientation of the input transformer in order to avoid hum pickup, especially if it is within a few feet of power transformers. Heavily-shielded input transformers of the "hum bucking" type are recommended for input transformers.

It is a good idea to design the amplifier for about 150 or 200 cycle cutoff, as this not only increases the effective modulation power (as explained in Chapter 14) but also minimizes hum troubles. This means that one can use inexpensive audio components, and also that one need not isolate the d.c. from the primary of a.f. transformers, because such isolation is required only for very low frequencies (below about 150 cycles). *Low harmonic distortion* is of more importance in getting a good sounding amateur signal than is wide-range frequency response.

The foregoing is more appropriately and extensively covered in the chapter on radiotelephony theory, but is mentioned here because it is so much tied in with transmitter design: how we lay out or plan the speech system of a transmitter depends upon just what features are to be incorporated and what requirements must be met. Before planning a speech amplifier or modulator one should read both the chapter on *Radiotelephony Theory* and the chapter on *Workshop Practice.*

Mains Supply. The problem of supplying the transmitter with alternating current power from the supply mains and turning the transmitter on and off, and "standby" while listening, is a problem that can be attacked in many ways, the "best" method being a matter of individual preference.

To make sure that an outlet will stand the full load of the entire transmitter, plug in an electric heater rated at about 50 per cent greater wattage than the power you expect to draw from the line. If the line voltage does not drop more than 5 volts (assuming a 117 volt line) under load and the wiring does not overheat, the wiring is adequate to supply the transmitter. About 750 watts total drain is the maximum that should be drawn from a 117 volt "lighting" outlet or circuit. For greater

power, a separate pair of heavy conductors should be run right from the meter box. For a 1 kw. phone transmitter the total drain is so great that a 220 volt "split" system ordinarily will be required. Most of the newer homes are wired with this system, as are homes utilizing electricity for cooking and heating.

With a 3-wire system, be sure there is no fuse in the neutral wire at the fuse box. A neutral fuse is not required if both "hot" legs are fused, and, should a neutral fuse blow, there is a chance that damage to the radio transmitter will result.

If you have a high power transmitter and do a lot of operating, it is a good idea to check on your local power rates if you are on a straight "lighting" rate. In some cities a lower rate can be obtained (but with a higher "minimum") if electrical equipment such as an electric heater drawing a specified amount of current is permanently wired in. It is not required that you use this equipment, and many an amateur who runs his kilowatt phone rig far into the night has made a worthwhile saving on his electric bill by scaring up an old 3 kw. air heater at the secondhand store and permanently installing it in the operating room. Naturally, however, there would be no saving unless you expect to occupy the same dwelling for a considerable length of time.

Probably the most popular transmitter switching system is the one shown in Figure 2. All transmitter tube filaments and possibly the speech amplifier plate voltage are turned on by means of one primary switch. With this switch on, the transmitter is in "standby" position (as soon as any mercury vapor rectifiers have once reached operating temperature).

Another switch, the "send-receive" switch S_2, is connected so as to control all plate transformers except possibly that used for the speech amplifier (which usually is a combined plate-filament transformer). This is perhaps the simplest method, but requires that the modulator and all r.f. tubes be supplied from filament windings that are not combined with plate windings on the same core. As this is common transformer practice anyway, except for low voltage supplies, no special requirements need be considered when purchasing transformers.

The send-receive switch in this system should be capable of handling the required power with considerable to spare, because of the inductive nature of the load. Thirty ampere mercury switches may be purchased for less than a dollar, and besides having a smooth and positive action, they will last almost indefinitely. They resemble an ordinary house lighting toggle switch in appearance. The latter, costing less than the mercury type, will

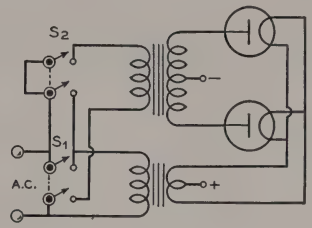


Figure 3.
FOOLPROOF RECTIFIER PROTECTION.
No matter which switch is thrown first, the filaments always will be turned on first and off last. The primaries of other filament transformers are connected in parallel with the primary of the rectifier filament transformer.

be found satisfactory in low-powered transmitters.

Another popular arrangement is to use fixed safety bias on the entire transmitter, so that the excitation may be removed at the "front end" of the transmitter without any of the succeeding tubes becoming overheated or going into parasitic oscillation. The transmitter then is turned on and off (or keyed, for that matter) simply by opening and closing the cathode or screen of the oscillator.

To minimize the external wiring, the most common practice is to turn the filaments on right at the transmitter, only the send-receive switch being placed on the operating desk, as in Figure 2. When the transmitter is small and is placed right on or beside the operating desk, both filament and send-receive switches may be placed on the transmitter.

In Figure 3 is shown an arrangement which protects mercury vapor rectifiers against premature application of plate voltage without resorting to a time delay relay. No matter which switch is thrown first, the filaments will be turned on first and off last. However, double pole switches are required in place of the usual single pole switches.

Safety Precautions

The best way for an operator to avoid serious accidents from the high voltage supplies of a transmitter is for him to use his head, act only with deliberation, and not take unnecessary chances. However, no one is infallible, and chances of an accident are greatly lessened if certain factors are taken into consideration in the design of a transmitter, in order to protect the operator in the event of a lapse of caution. If there are too many things one must "watch out for" or keep in mind there is a

good chance that sooner or later there will be a mishap; and it only takes *one*. When designing or constructing a transmitter, the following safety considerations should be given attention.

Grounds. For the utmost in protection, everything of metal on the front panel of a transmitter capable of being touched by the operator should be at ground potential. This includes dial set screws, meter "zero adjuster" screws, meter cases if of metal, meter jacks, *everything* of metal protruding through the front panel or capable of being touched or *nearly* touched by the operator. This applies whether or not the panel itself is of metal. Do not rely upon the insulation of meter cases or tuning knobs for protection.

The B negative or chassis of all plate power supplies should be connected together, and to an external ground such as a waterpipe. In the case of a bias supply, the B positive should be connected to the common ground.

Exposed Wires and Components. It is not necessary to resort to rack and panel construction in order to provide complete enclosure of all components and wiring of the transmitter. Even with breadboard construction it is possible to arrange things so as to incorporate a protective housing which will not interfere with ventilation yet will prevent contact with all wires and components carrying high voltage d.c. or a.c.

If everything on the front panel is at ground potential (with respect to external ground) and all units are effectively housed with protective covers, then there is no danger except when the operator must reach into the interior part of the transmitter, as when changing coils, neutralizing, adjusting coupling, or shooting trouble. The latter procedure can be made safe by making it possible for the operator to be *absolutely certain* that all voltages have been turned off and that they cannot be turned on either by short circuit or accident. This can be done by incorporation of the following system of main primary switch and safety signal light.

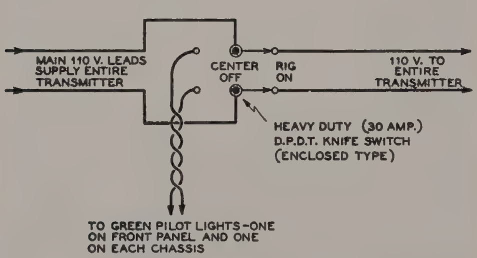


Figure 4.

COMBINED MAIN SWITCH AND SAFETY SIGNAL.

After shutting down the transmitter for the day, throw the main switch to neutral. If you are going to work on the transmitter, throw the switch all the way to "pilot," thus turning on the green pilot lights and making it impossible for there to be primary voltage on any transformer in the transmitter even by virtue of a short or accidental ground. To live to a ripe old age, simply obey the rule of "never work on the transmitter unless green lights are on."

Combined Safety Signal and Switch. The common method of using red pilot lights to show when a circuit is "on" is useless except from an ornamental standpoint. When the red pilot is not lit it *usually* means that the circuit is turned off, but it *can* mean that the circuit is on but the lamp is burned out or not making contact.

To enable you to grab the tank coils in your transmitter with absolute assurance that it is impossible for you to obtain a shock except from possible undischarged filter condensers (see following topic for elimination of this hazard), it is only necessary to incorporate a device similar to that of Figure 4. It is placed near the point where the main 110-volt leads enter the room (preferably near the door) and in such a position as to be inaccessible to small children. Notice that this switch breaks *both* leads; switches that open just one lead do not afford complete protection, as it is sometimes possible to complete a primary circuit through a short or accidental ground. Breaking just one side of the line may be all right for turning the transmitter on and off, but when you are going to stick an arm inside the transmitter, *both* 110-volt leads should be broken.

When you are all through working your transmitter for the time being, simply throw the main switch to neutral. Then you can leave the transmitter and even go on a vacation with absolute peace of mind.

When you find it necessary to work on the transmitter or change coils, throw the switch so that the green pilots light up. These can be ordinary 15-watt green bulbs. One should be placed on the front panel of the transmitter; others should be placed so as to be easily visible when changing coils or making adjustments requiring the operator to reach inside the transmitter. These lamps are inexpensive, and as several will draw less than 100 watts from the line, a half dozen may be scattered around the transmitter.

For 100 per cent protection, just obey the following rule: *never work on the transmitter or reach inside any protective cover except when the green pilots are glowing.* To avoid

confusion, no other green pilots should be used on the transmitter; if you want an indicator jewel to show when the filaments are lit, use amber instead of green.

If the main switch is out of reach of small children, a conspicuous sign, such as "DO NOT TOUCH UNDER ANY CIRCUMSTANCES," placed on the switch cover will guard against the off chance that someone else would throw the switch unexpectedly. An alternative is to place the switch on the under side of the operating table out of sight. The latter is not so desirable when small children have access to the room.

Safety Bleeders. High capacity filter condensers of good quality hold their charge for some time, and when the voltage is more than 1000 volts it is just about as dangerous to get across an undischarged 4-μfd. filter condenser as it is to get across a high voltage supply that is turned on. Most power supplies incorporate bleeders to improve regulation, but as these are generally wire-wound resistors, and as wire-wound resistors occasionally open up without apparent cause, it is desirable to incorporate an auxiliary safety bleeder across each heavy-duty bleeder. Carbon resistors will not stand much dissipation and sometimes change in value slightly with age. However, the chance of their opening up when run well within their dissipation rating is almost unheard of.

To make *sure* that all condensers are bled, it is best to short each one with an insulated screwdriver. However, this is sometimes awkward and always inconvenient. One can be virtually sure by connecting auxiliary carbon bleeders across all wire-wound bleeders used on supplies of 1000 volts or more. For every 500 volts, connect in series a 500,000-ohm 1-watt carbon resistor. The drain will be negligible (1 ma.) and each resistor will have to dissipate only 0.5 watt. Under these conditions the resistors will last indefinitely with no chance of opening up. For a 1500-volt supply, connect three 500,000-ohm resistors in series. If the voltage exceeds an integral number of 500 volt divisions, assume it is the next higher integral value; for instance, assume 1800 volts as 2000 volts and use four resistors.

Do *not* attempt to use fewer resistors by using a higher value for the resistors; not over 500 volts should appear across any single 1-watt resistor.

In the event that the regular bleeder blows, it will take several seconds for the auxiliary bleeder to drain the condensers down to a safe voltage, because of the very high resistance. Hence, it is best to allow 10 or 15 seconds after turning off the plate supply before attempting to work on the transmitter.

"Hot" Adjustments. Some amateurs contend that it is almost impossible to make certain adjustments, such as coupling and neutralizing, unless the transmitter is running. The best thing to do is to make all neutralizing and coupling devices adjustable from the front panel by means of flexible control shafts which are broken with insulated couplings to permit grounding of the panel bearing.

If your particular transmitter layout is such that this is impracticable and you refuse to throw the main switch to make an adjustment —throw the main switch—take a reading—throw the main switch—make an adjustment —and so on, then protect yourself by making use of long adjusting rods made from ½-inch dowel sticks which have been wiped with oil when perfectly free from moisture.

If you are addicted to the use of pickup loop and flashlight bulb as a resonance and neutralizing indicator, then fasten it to the end of a long dowel stick and use it in that manner.

CHAPTER TWELVE

Exciters and Low Powered Transmitters

SIMPLE 15-WATT TWO BAND EXCITER OR TRANSMITTER

Illustrated in Figures 1 and 2 is the simplest practical exciter or transmitter for fixed-station use. It uses only one tube and one crystal, and with four easily wound coils provides about 15 watts output on 80 meters and approximately 12 watts on 40. With few exceptions, the parts are all inexpensive standard receiver items. With the particular antenna-coupling circuit illustrated, the unit may be used with a wide variety of antennas, although the simple antenna to be described is strongly recommended. It gives excellent performance on both bands.

The unit operates as a regenerative crystal oscillator of the harmonic type on 40 meters and as a straight tetrode crystal oscillator on 80 meters. The change from one form of oscillator to the other is taken care of automatically when the coils are changed, as a result of the jumper in the 80-meter coil.

If the unit is used as an exciter, the antenna coupling tank L_2 and C_6 may be omitted, the output of the oscillator being link coupled to the following stage instead. The antenna tank circuit illustrated was included in the model shown because it can be used in conjunction with an end-fed wire for 2-band operation. If the unit is first used as a transmitter and then later used as an exciter when another stage is added, the antenna tank circuit can be removed from the oscillator unit and used as the grid tank of the amplifier.

Construction. The whole transmitter is built on a $9\frac{1}{2}$ x $6\frac{1}{2}$ x 1 inch thick wooden baseboard to which is mounted a $10\frac{1}{2}$ x $6\frac{1}{2}$ inch "Presdwood" front panel.

Baseboard-mounting type bakelite sockets are used for both the tube and the coils. Five-prong sockets are used for the coils and a 6-prong one for the tube. Another 5-prong socket of the same type is placed directly behind the tube and used to mount the crystal.

The panel supports the two midget "tank" condensers, C_1 and C_6, and the 0-100 ma. meter. A small through-type insulator directly above the antenna-tuning condenser is used for an antenna terminal.

The two Fahnestock clips at the right rear of the baseboard are used for key connections. A small 4-terminal strip at the left rear of the baseboard provides a convenient method of making heater and plate voltage connections to the power supply. The only other components mounted on either the panel or base are two 2-terminal tie points. These are screwed to the baseboard, one between each coil and condenser. They are used to support the coupling links, which will be described later.

Wiring. With the exception of the coupling link, the heater leads, and one of the meter leads, all wiring is done with no. 14 busbar. This heavy wire allows the various fixed condensers and resistors to be supported directly from the wiring.

A single piece of bus-bar running along the back of the baseboard between the tube and the crystal, and connected to one of the power supply terminals at one end, and to one of the key terminals at the other is used for a *common ground* lead. All of the ground connections shown on the diagram are made to this lead, which in turn should be connected to a waterpipe or other good external ground.

As may be seen from the diagram, there is a link around each coil. These links couple the plate coil to the simple antenna-matching circuit. The link around the plate coil is 3 turns of push-back wire, while the one around the antenna coil is 4 turns of the same type of wire. The links are each $1\frac{3}{4}$ inches in diameter and are permanently connected in the transmitter. They are supported by the tie-points previously mentioned. Two small pieces of

tape wrapped around each link coil serve to hold the turns together. The link around the plate coil should be placed at such a height above the socket that when the plate coil is plugged in, the link is around the bottom portion of the coil. The bottom of the plate coil should be the end which is connected through C_3 to ground on 40 meters and, by means of the jumper, directly to ground on 80 meters.

The link coil around the antenna coil should be positioned so that it falls at the *center* of the *antenna coil*. About 6 inches of twisted push-back wire is used as a coupling line between the two coils. The twisted line is connected to tie-points at each end of the line.

Coils. The jumper on the 80-meter coil allows the transmitter to work as a conventional tetrode oscillator on 80 meters and as a regenerative oscillator on 40 meters.

The antenna coil connections are the same for both bands. If the socket connections are made as shown in the diagram, the two ends of the coils are connected to the cathode and plate prongs, and the center tap to the grid prong.

The leads to the key may be any reasonable length (up to 10 feet, if necessary). A 0.02-μfd. condenser, C_7, is connected directly *across the key*. This condenser is used to minimize key clicks and is most effective when placed right at the key rather than in the transmitter. Be sure the *frame* of the key connects to the grounded key terminal and not the terminal that goes to the meter.

Power Supply. The power supply recommended is a standard brute-force filtered affair using receiver components throughout. The parts are mounted on a small baseboard in the most convenient manner, and the heater and plate voltage connections brought out to a 4-post terminal strip similar to that on the transmitter. The power transformer should not deliver more than 350 volts r.m.s. each side of the c.t. or else the peak voltage on the filter condensers will be too high when the key is up.

Antenna. The best type of antenna for use with this transmitter is the end-fed half-wave 80-meter type. Such an antenna, if erected reasonably in the clear, will give good results on both 80 and 40 meters. On both bands the antenna is not particularly directional, although a slight increase in signal strength will be noticed in certain directions. On 40 meters the antenna produces low-angle

COIL TABLE

Band	Plate Coil	Antenna Coil
80	41 turns, close-wound	50 turns, center-tapped, close-wound
40	21 turns, spaced to a length of two inches	26 turns, center-tapped, spaced to a length of two inches

All coils wound with no. 20 double-cotton-covered wire on 1½″ dia. forms.

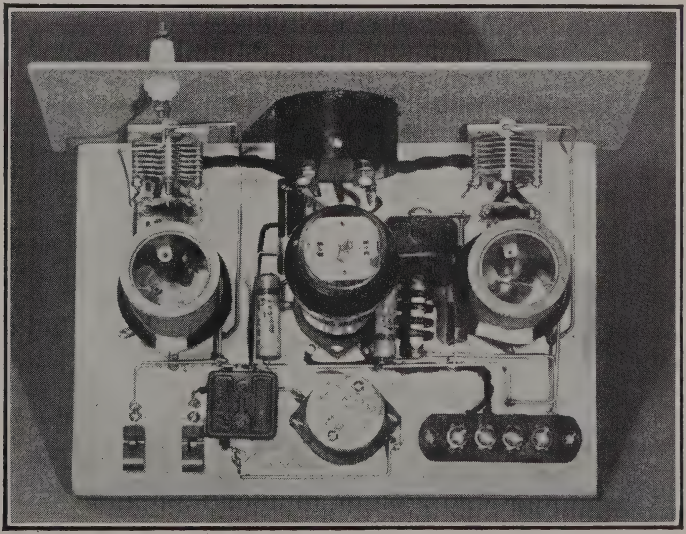


Figure 1.
SIMPLEST PRACTICAL EXCITER OR TRANSMITTER.
This unit delivers 12 to 15 watts on 40 or 80 meters with an 80-meter crystal. The antenna coupling tank (to the left) can be omitted if this unit is to be used only as an exciter.

radiation, an advantage in working dx.

The antenna should measure 135 feet from the far end *to the antenna terminal on the transmitter,* and be erected in the clear and as high and as much in a straight line as possible.

Tuning Up. If all the wiring has been done properly, no difficulty should be experienced in placing the transmitter in operation. Leads to the power supply and key should be connected (ordinary lamp cord of good quality will do), and a 6.3-volt 150-ma. dial light placed in series with the antenna at the transmitter. A crystal with a frequency between 3502 and 3648 kc. should be placed in the crystal socket. A crystal in this range will allow operation on both the 80- and 40-meter bands.

When the transmitter is properly adjusted for 80-meter operation, it should be possible to tune the antenna coupling circuit through resonance without pulling the oscillator out of oscillation. The dial light should increase in brilliance as the antenna circuit is tuned up to resonance and then decrease as it is detuned from resonance on the other side.

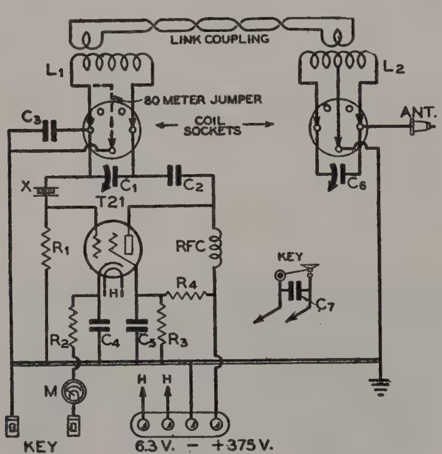


Figure 2.
THE R.F. PORTION OF THE TRANSMITTER.

C_1 — 50-$\mu\mu$fd. midget variable
C_2—.01-μfd. mica
C_3—.0005-μfd. mica
C_4, C_5 — .01-μfd. 600-volt tubular
C_6 — 50-$\mu\mu$fd. midget variable
C_7 — .02-μfd. 600-volt tubular
R_1—100,000 ohms, 1 watt
R_2 — 400 ohms, 10 watts
R_3—20,000 ohms, 10 watts
R_4 — 5000 ohms, 10 watts
RFC — 2.5-mh., 125-ma. choke
X—80-meter X or AT crystal
L_1, L_2—See coil table
M — 0-100 milliamperes

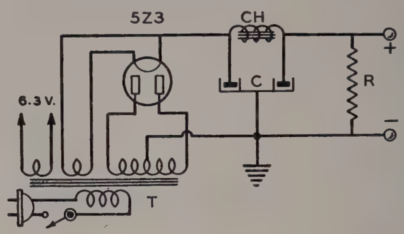


Figure 3.
RECOMMENDED POWER SUPPLY.

T—700 v.c.t., 90 ma.; 5 v., 3 a.; 6.3 v., 3 a.
CH—30 hy., 110 ma.
C — Dual 8-μfd. electrolytic, 450 v.
R — 40,000 ohms, 20 watts

When this condition is obtained, remove the dial lamp from the antenna and make the antenna connection directly to the antenna post. Then, without touching the antenna-tuning condenser, turn the plate condenser toward maximum capacity until the point of maximum capacity at which the circuit will still oscillate is found. The final adjustment of the plate condenser should be made while listening to the signal from the transmitter in a monitor or receiver. The condenser should be set at the furthest point toward maximum capacity at which the keying is clean and distinct without chirps or lag.

The farther down the plate coil the coupling link is placed, the looser the coupling to the antenna circuit. If the coupling is too tight, the oscillator won't oscillate or the note will be chirpy. If the coupling is too loose, the full power will not be delivered to the antenna.

The coupling should be adjusted by varying the position of the *plate coil coupling link,* never by detuning the antenna condenser. The latter should always be tuned to resonance. If it cannot be tuned to resonance without the transmitter's going out of oscillation or developing keying chirps, the coupling is too tight.

If the dial lamp in the antenna lead does not give sufficient indication to be observed handily, a 2-volt 60-ma. bulb may be substituted. Do not use a 60-ma. lamp unless you are unable to get a satisfactory indication on a 150-ma. bulb. The maximum antenna current will be low at this point (a current "node") and will vary somewhat in different antenna installations.

On 40 meters the tuning is somewhat simpler, because the transmitter acts as a regenerative harmonic oscillator and will oscillate and key cleanly regardless of how

heavily the plate circuit is loaded. Therefore, it is necessary only to tune for greatest output, without regard to keying chirps or nonoscillation.

When the unit is used as an exciter the tuning is the same except that instead of tuning for greatest brilliancy of the lamp in the antenna lead, adjustments should be made for maximum grid current to the following stage. Coupling is adjusted as described for operation with an antenna; the position of the link around L_1 is varied until the desired coupling is obtained. *Be sure to turn off the power supply* before making coupling adjustments.

5-WATT 160 METER V.F.O.

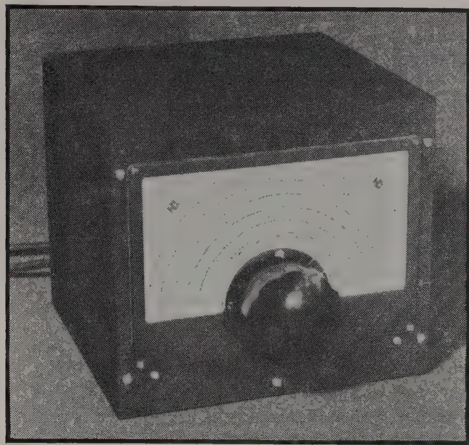


Figure 4.
5-WATT 160 METER V.F.O. OF HIGH STABILITY.
Delivering its output in the 160 meter band, this unit can be used for all band operation. It takes up very little room on the operating table, the frequency doublers being made part of the transmitter proper instead of being incorporated in the frequency control unit.

Illustrated in Figures 4, 5, 6, and 7 is a variable frequency exciter which is very stable, is free from drift, and uses relatively few parts. The output is on 160 meters, which permits operation on all bands. To minimize the size of the unit on the operating table, frequency doublers are incorporated in the transmitter proper instead of being made an integral part of the v.f.o. unit. The output is approximately 5 watts.

To minimize frequency drift from heating, the two amplifier stages are allowed to run continuously, the oscillator being switched on and off with the main transmitter. Thus, the amplifier stages dissipate approximately the same amount of heat regardless of whether or not the oscillator is running. These two stages derive plate supply from a small 275-volt pack which is turned on with the transmitter filaments; it feeds no other stages.

To reduce drift further, this power pack and the voltage regulator tube for the oscillator are made external to the unit, and all tubes in the v.f.o. unit are mounted so as to project from

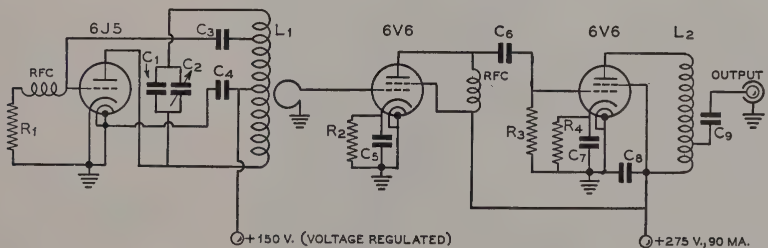


Figure 5.
SCHEMATIC DIAGRAM OF 160 METER V.F.O.

C_1—335-$\mu\mu$fd. midget condenser (used as fixed padder)
C_2—150-$\mu\mu$fd. midget condenser, double bearing type, semi-circular plates
C_3—100-$\mu\mu$fd. zero temperature coefficient mica condenser
C_4—.006-μfd. mica
C_5—.05μfd. tubular
C_6—.006-μfd. mica
C_7, C_8—.05-μfd. tubular
C_9—.006-μfd. mica
R_1—50,000 ohms, ½ watt
R_2—300 ohms, 1 watt
R_3—1000 ohms, 1 watt (must be carbon)
R_4—300 ohms, 1 watt
RFC—2.5 mh. choke
L_1, L_2—Refer to text

the cabinet as shown in the photographs. The greatest portion of the heat radiated by the tubes is kept outside the cabinet, and the remainder is kept from affecting the oscillator components by proper ventilation of the cabinet. Except for a few minutes when the device is first turned on, the drift is negligible even when doubling to 10 meters.

A standard high-C Hartley oscillator feeds an untuned class A buffer, which in turn feeds a fixed-tuned class A amplifier. This arrangement provides excellent isolation of the oscillator, yet requires only one tuning condenser.

The output circuit is designed for use with a 70-ohm coaxial line to the transmitter proper. This may be of the inexpensive, flexible type having rubber dielectric. If the following tube requires 25 peak volts or less of r.f. voltage, no tuned circuit is required at the far end of the line. The line simply is terminated at the far end in a resistor consisting of two 150-ohm 2-watt carbon resistors in parallel, giving a terminating resistance of 75 ohms. The grid of the tube to be excited is coupled through a blocking condenser from the "hot" end of the terminating resistor. A 6L6, 6V6, 807, etc. driven in this manner will receive sufficient excitation for efficient operation either as a doubler or straight amplifier, and the coupling line may be made any length without detrimental effects.

Where greater drive is required, such as might be the case where a tube such as an 813, 809, etc. is to be excited, the coaxial line is link-coupled to a tuned circuit at the transmitter end in the conventional manner. Using this arrangement, the full output of the exciter will be available as excitation to the transmitter. Thus, much greater driving voltage is obtained, but at the expense of an additional tuned circuit.

All parts of the oscillatory circuit are made as solid mechanically as is possible. The stability of a v.f.o. depends as much as anything upon mechanical construction. While a cast type chassis and cabinet would be preferable, the standard type shown is satisfactory because of its small size. A larger cabinet and chassis of this type would lack sufficient rigidity.

The cabinet measures 7 inches high by 8 inches wide by 8 inches deep. The chassis measures 7 inches by 7 inches by $1\frac{1}{2}$ inches deep. For the sake of rigidity, a "closed end" type chassis should be chosen in preference to an open ended type. Holes are drilled in the rear of the cabinet to accommodate not only the power and output cables, but also the three tubes, as shown in Figure 7. Several $\frac{3}{8}$-inch holes are drilled along the top rear of the cabinet, as shown in this illustration, to encourage what heat is generated or transferred inside the cabinet to be carried out as quickly as possible by convection.

All oscillator components are placed above chassis, all amplifier components below. This minimizes capacity coupling feedback, as the only r.f. connection between the upper and

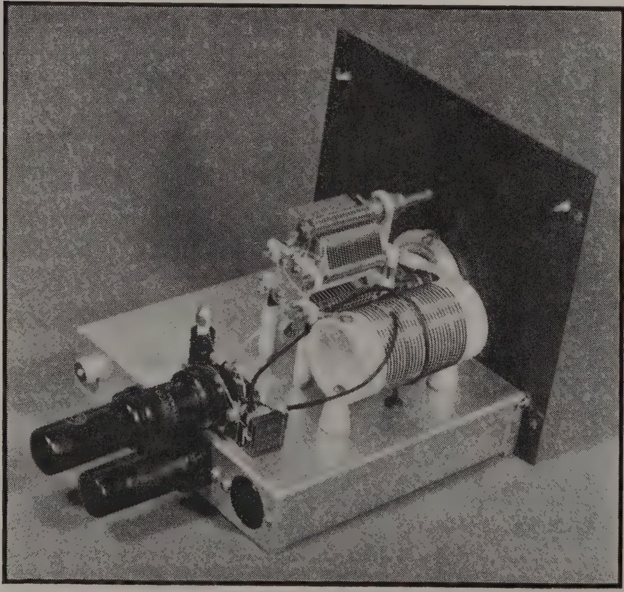


Figure 6.
INTERIOR CONSTRUCTION OF 160 METER V.F.O.
All parts of the oscillator circuit are made very rigid. The connector at the far end of the chassis back-drop is for the coaxial output cable, the near one for power connections.

lower decks is via the single turn coupling link around the cold portion of the oscillator coil.

The socket for the oscillator tube is mounted by means of heavy brackets so that the tube may be allowed to protrude out the rear of the cabinet in the same manner as the amplifier tubes. The tuning condenser is supported by means of 1½-inch ceramic pillars and small brackets. The padding condenser is supported on the tuning condenser by means of small strips of aluminum cut and drilled to accommodate the condenser bolts and studs so that the two condensers are mounted back to back about ½-inch apart. The tuning condenser is mounted upside down, as is apparent in the illustration. The assembly is further strengthened by an aluminum cross brace between the stator lug of the top condenser and one end of the coil form. This brace also serves as an electrical connection between the condenser and coil. The ceramic coil form is raised off the chassis by means of ¾-inch ceramic pillar insulators.

As the rotor of the tuning condenser is "hot," it must be insulated from the metal parts of the dial (which are grounded). This automatically is taken care of in the particular dial used in the unit illustrated, as an insulated coupling is an integral part of the dial mechanism. A smooth working dial with no backlash is a virtual necessity.

Because of the comparatively large amount of lumped padding capacity across the tank, a tuning condenser with semi-circular plates will provide a more uniform distribution of kilocycles than will a modified plate shape. The frequency coverage is from 1750 to 2050 kc., with a few kilocycles overlap at each end.

Coils. The temperature coefficient of the oscillator coil is minimized by using a ceramic form and winding on the wire in such a manner that it is under tension. The wire first is stretched until it is near the breaking point. Then, with both wire and form maintained at a temperature as high as will permit handling, the wire is wound tightly on the form. When the wire and form cool to room temperature, the wire will be under considerable tension, and the linear expansion of the wire with changes in room temperature will be greatly reduced. The temperature coefficient of the ceramic form is only about ⅕ that of the copper wire, while the coefficient of a bakelite form (commonly used for the purpose) is about twice that of the copper wire. Another advantage of the ceramic form is that its white color reflects any small amount of radiant heat that may reach it, whereas black bakelite would not. Because of these various precautions, no temperature compensating capacitors are required in order to avoid frequency drift.

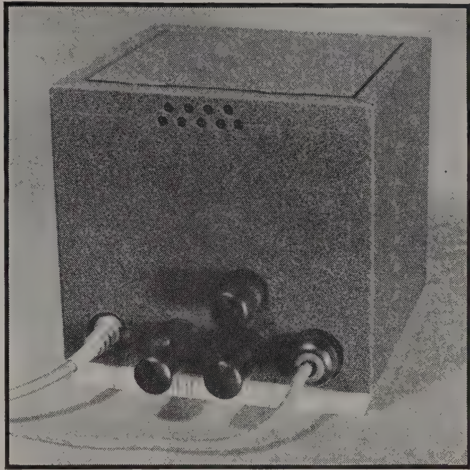


Figure 7.
REAR VIEW OF 160 METER V.F.O. CABINET.

Holes are punched in the rear of the cabinet just large enough to take the tubes. Air entering the lower side louvres and leaving by the holes in the top rear of the cabinet keeps what little heat is generated inside the cabinet from affecting the oscillator components.

The ceramic form measures 1¾ inches in diameter by 3½ inches long. The coil contains 26 turns of no. 20 enamelled, spaced to cover 2⅝ inches on the form. The wire first is stretched as previously described and wound on the form while warm. The coil is tapped at the exact center and at 6 turns from the center. In making the tapped connections, care must be taken not to heat the wire on the form any more than is necessary to make a good soldered joint, or it will lose its tension on the form. If this happens, the coil should be heated and the wire pulled tight again.

The untuned output coil consists of 82 turns of no. 26 d.c.c. scramble-wound over a space of about ½ inch on a 1-inch diameter bakelite form. It is tapped 8 turns from the ground end for the output connection. The turns are held in place by coil dope or Duco cement.

The coupling link from the oscillator to the buffer consists of a single turn of hookup wire around the center of the coil.

Operation. The unit should be set on a piece of sponge rubber to protect it from jars or vibration. A piece of typewriter pad or "kneeling pad" will be satisfactory. When possible, the exciter should be allowed to warm up for 5 or 10 minutes before operating the v.f.o. unit on the air.

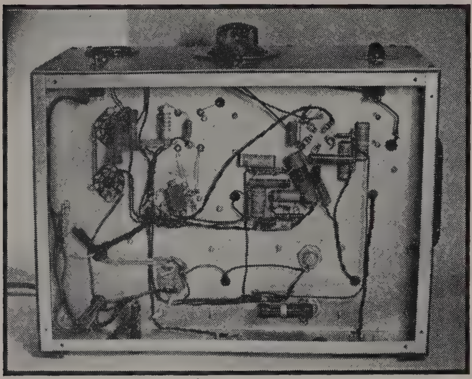


Figure 8.
UNDER-CHASSIS VIEW OF 25-WATT V.F.O.

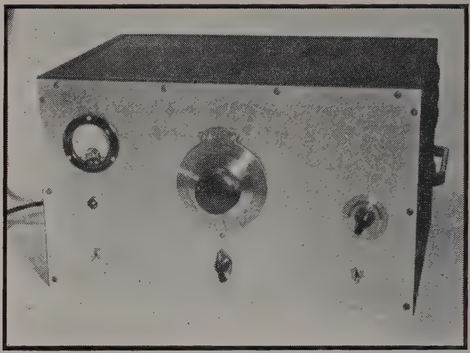


Figure 9.
20-25 WATT V.F.O. FOR 20, 40, AND 80 METERS.
Incorporating an 807 output tube and integral 500-volt power supply, this v.f.o. is capable of considerably more output than most. Because of the high power output and multi-band operation, it is more properly termed a "v.f.o.-exciter."

The cascaded frequency multiplier unit described later in this chapter is particularly well suited for use with this v.f.o. unit, providing output on 160, 80, 40, or 20 meters at the flip of a switch.

25-WATT V.F.O. FOR 80, 40, AND 20 METERS

Illustrated in Figures 8–11 is a v.f.o. unit which delivers 20–25 watts on either 80, 40, or 20 meters simply by changing one coil. The power supply is self-contained, drift as a result of heat being minimized by proper compensation. The unit is patterned after a design developed by G. W. Shuart.

The 6SJ7 electron-coupled oscillator operates on 160 meters, covering the range from 1750 to 2000 kc. The frequency control components are procurable as a stardard manufactured unit which has been designed for mechanical stability and low temperature coefficient of frequency. The unit is comprised of a shield can which houses L_1, C_3, C_4, C_6, C_8, and R_1.

The output of the 6SJ7 oscillator is tuned to 80 meters by a broadly tuned tank, also avail-

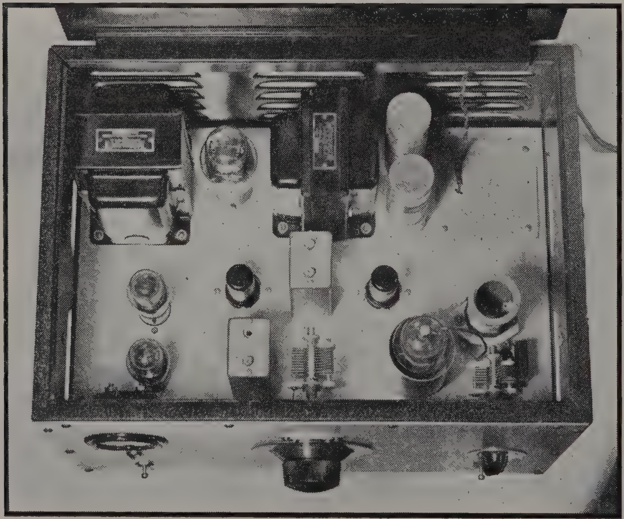


Figure 10.
INTERIOR CONSTRUCTION OF V.F.O.-EXCITER.
Because the required characteristic of the temperature compensating condenser is determined largely by the physical construction of the unit, it is recommended that the layout shown here be followed closely.

able as a standard unit. The characteristics of this unit are such that most of the 3500–4000 kc. band can be covered without the necessity for retuning the trimmer condenser. When operating on higher frequency bands, the trimmer is set for about 3600 kc. and left alone.

The 80-meter tank excites a 6SK7 whose output is fixed-tuned to a frequency near the center of the 7 Mc. band. The permeability tuned 40-meter coil is a stock item. The following stage, the 807, is tuned either to 20 meters (on which frequency it operates as a doubler), to 40 meters, or to 80 meters. Enough 80-meter excitation is developed across the low-C 40-meter tank L_3 to permit the 807 to operate on 80 meters with about the same efficiency as is obtained when doubling. Thus, to obtain output on any of the three bands, it is necessary only to change the output coil, L_4.

On the higher frequency bands it will be

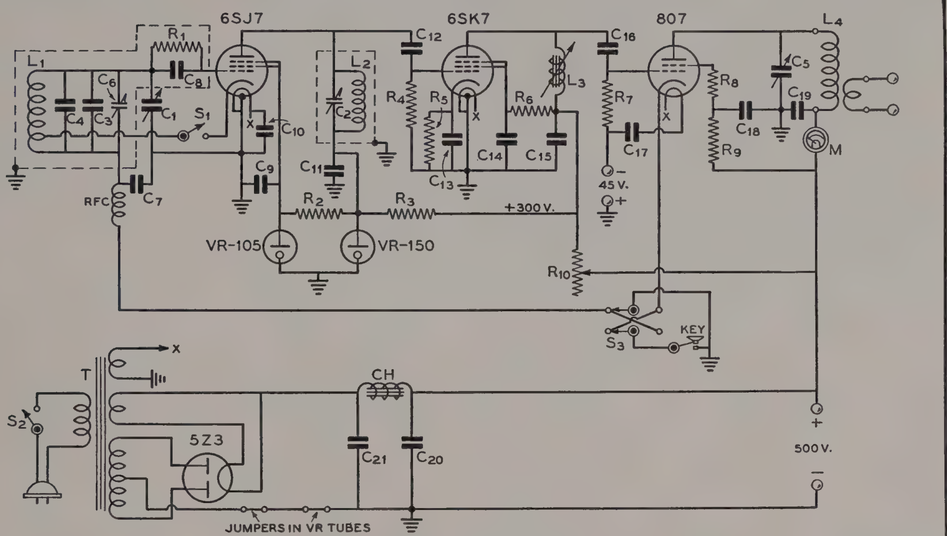


Figure 11.

SCHEMATIC DIAGRAM OF 25-WATT V.F.O. UNIT.

C_1 — 140-μμfd. midget condenser, straight line capacity type, with rigid plates and good bearings

C_2 — 50-μμfd. air trimmer (part of manufactured tank assembly)

C_3 — 275-μμfd. silvered mica fixed condenser, zero temperature coefficient type (part of manufactured tank assembly)

C_4—35-μμfd. negative coefficient condenser (part of manufactured tank assembly)

C_5 — 100-μμfd. midget condenser

C_6 — 75-μμfd. air trimmer (part of manufactured tank assembly)

C_7—.005-μfd. mica

C_8 — 100-μμfd. silvered mica condenser (part of manufactured assembly)

C_9—.01-μfd. tubular

C_{10} — .001-μfd. mica, right at osc. tube terminals

C_{11}—.01-μfd. tubular

C_{12} — .0005-μfd. mica

C_{13}, C_{14}, C_{15} — .01-μfd. tubular

C_{16} — .0005-μfd. mica

C_{17}, C_{18} — .01-μfd. tubular

C_{19} — .005-μfd. mica, 1000 v. test

C_{20}—4-μfd., 600 v. working

C_{21}—2-μfd., 1000 v. working

R_1 — 20,000 ohms, ½ watt

R_2, R_3—3000 ohms, 10 watts

R_4 — 20,000 ohms, 1 watt

R_5 — 400 ohms, 1 watt

R_6 — 50,000 ohms, 1 watt

R_7 — 20,000 ohms, 1 watt

R_8 — 25 ohms, 1 watt carbon

R_9 — 20,000 ohms, 10 watts

R_{10} — 5000 ohms, slider type, 50 watts (adjust to give 300 v. to plate of 6SK7)

RFC — 2.1 mh. choke

S_1 — Standby switch, low capacity. Must have short leads.

S_2 — 110 v. a.c. switch

S_3—D.p.d.t. switch for switching key from osc. to amp.

M—0-100 or 0-150 ma. d.c.

L_1, L_2, L_3, L_4—Refer to text.

T — 870 v. center tapped, 250 ma., with 5 v. and 6.3 v. fil. windings

CH—13 hy. at 250 ma.

unnecessary to retune L_4 when moving from one portion of the band to another, but to cover the extreme limits of the 80-meter band it will be necessary to readjust C_5 for maximum output.

Compound voltage regulation is used on the oscillator screen, and standard regulation on the plate. This makes the oscillator absolutely immune to changes in line voltage.

By means of S_3, it is possible to change from oscillator keying to amplifier keying simply by throwing a switch. Fixed battery bias on the 807 reduces the plate and screen current to this stage to a very low value when the key is up during oscillator keying. By adjusting the switch S_3 for amplifier keying, it is possible to set the frequency with the aid of a receiver without putting a signal on the air simply by shutting the standby switch S_1 while the key is open. This assumes that the balance of the transmitter uses safety bias or else that the plate voltage is removed for the moment.

Coils. As mentioned previously, all coils except L_4 are available as standard items. However, for those who might wish to construct their own, the specifications are given.

Coil L_1 consists of 30 turns of no. 24 enamelled, close-wound on a $\frac{7}{8}$-inch diameter ceramic form, with cathode tap 10 turns from the ground end, mounted in coil shield.

Coil L_2 consists of 60 turns of no. 24 enamelled wire, close-wound on $\frac{7}{8}$-inch diameter form.

Coil L_3 consists of 36 turns of no. 28, close-wound on $\frac{5}{8}$-inch diameter form with adjustable "tuning plug."

Coils for L_4 all are wound on standard $1\frac{1}{2}$-inch diameter 5-prong forms. The 80-meter coil consists of 38 turns of no. 24 spaced to $1\frac{5}{8}$ inches with a 10-turn link at the cold end.

The 40-meter coil consists of 18 turns of no. 20, spaced to $1\frac{1}{2}$ inches, with a 5-turn link at the cold end. The 20-meter coil consists of 9 turns of no. 16, spaced to $1\frac{1}{4}$ inches, with a 3-turn link at the cold end.

Because the requirements of the temperature compensating condenser are determined largely by the physical construction of the v.f.o. unit, it is important that the layout shown in the illustration be adhered to closely in order to keep the drift to a negligible value.

Data on the manufactured tank circuits will be found in the Buyer's Guide.

MULTI-BAND FREQUENCY MULTIPLIER

The cascaded doubler unit illustrated in Figures 12–15 will permit output on any four consecutive amateur bands between 5 and 160 meters at the throw of a switch. The only difference is in the coil specifications, which are given later on.

Output on the lowest frequency band will be determined by the power of the unit employed to excite the frequency multiplier, as on this band the excitation merely is "shunted around" the unit by the selector switch. Output of the doubler stages will be about 15 watts except on 5 meters, where the output will be close to 10 watts.

The unit consists of three 6L6 doubler stages, with provision for coupling out of any of them by means of a switch in the link circuit. This same switch also applies full screen voltage to the particular 6L6 which happens to be serving as the output tube, the others running at reduced screen voltage.

Except for the switching arrangement, the circuit is perfectly straightforward, excepting

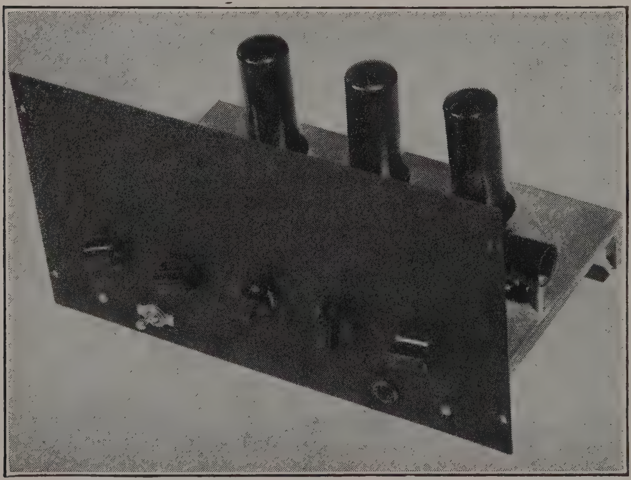


Figure 12.
MULTI-BAND FREQUENCY MULTIPLIER.
This unit delivers approximately 15 watts on 10, 20, or 40 meters when supplied with 80-meter excitation. No coils need be changed when changing from band to band.

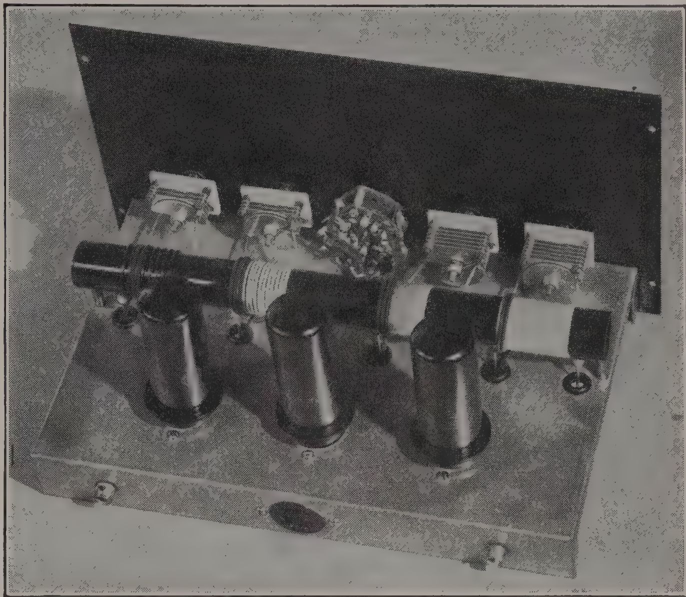


Figure 13.
SHOWING CONSTRUCTION OF MULTI-BAND FREQUENCY MULTIPLIER.
Because the coils all are tuned to a different band, there is little coupling between them even though mounted on a single winding form. Those shown here are for 80, 40, 20, and 10 meters.

possibly for the incorporation of fixed battery bias. A standard duty 45-volt bias battery costs no more than cathode resistors and bypass condensers for the three stages, will last for at least 2 years, and permits better oscillator keying.

Construction. The construction is shown clearly in the illustrations. A 7 x 11 inch chassis supports a 7 x 12 inch front panel. As the tank coils are all tuned to different bands, there is no need for taking special precautions to avoid coupling between coils. All four coils

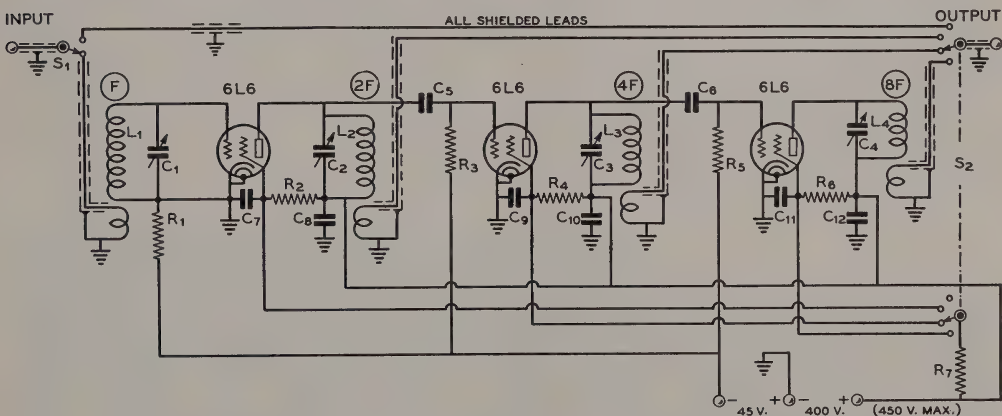


Figure 14.
SCHEMATIC DIAGRAM OF MULTI-BAND FREQUENCY MULTIPLIER.

C_1, C_2 — 35-$\mu\mu$fd. midget variable
C_3, C_4 — 25-$\mu\mu$fd. midget variable
C_5, C_6 — 25-$\mu\mu$fd. midget mica fixed
C_7 to C_{12}—.003-μfd. mica
R_1—100,000 ohms, 2 watts
R_2 — 50,000 ohms, 2 watts
R_3—100,000 ohms, 2 watts
R_4 — 50,000 ohms, 2 watts
R_5—100,000 ohms, 2 watts
R_6 — 50,000 ohms, 2 watts
R_7 — 15,000 ohms, 10 watts
S_1 — Single - pole double - throw toggle switch
S_2—2-pole 4-throw rotary switch
Coils—See text

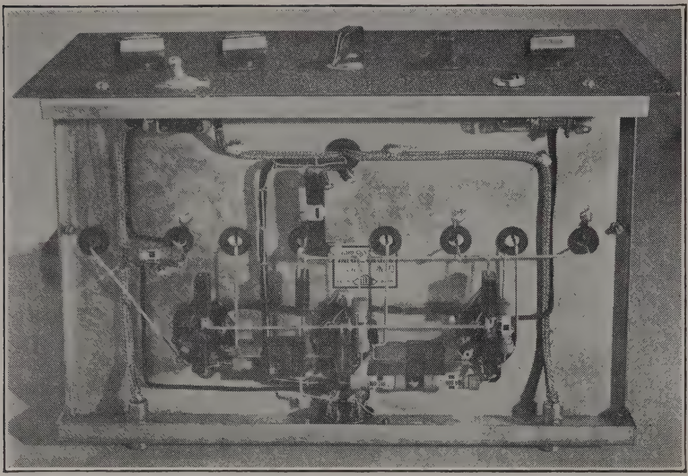


Figure 15.
UNDER-CHASSIS VIEW OF MULTI-BAND FREQUENCY MULTIPLIER.
Auto radio antenna couplers are used for making link connections in and out of the unit. Either shielded solid conductor or twisted hookup wire may be used for the link coupling line.

are wound on a single 10½-inch length of 1 inch dia. bakelite tubing. If desired, the coils can just as well be wound on four separate forms. Tuning condensers are supported from the chassis (not the panel) by means of brackets furnished by the manufacturer.

Coils. The 160-meter coil consists of 120 turns of no. 28 enamelled wire, close-wound, with a 4-turn link at the ground end.

The 80-meter coil consists of 48 turns of no. 24 d.c.c. close-wound, with a 3-turn link at the ground end.

The 40-meter coil consists of 23 turns of no. 20 d.c.c., close-wound, with a 3-turn link at the ground end.

The 20-meter coil consists of 13 turns of no. 18 d.c.c., spaced to 1 inch, with a 3-turn link at the ground end.

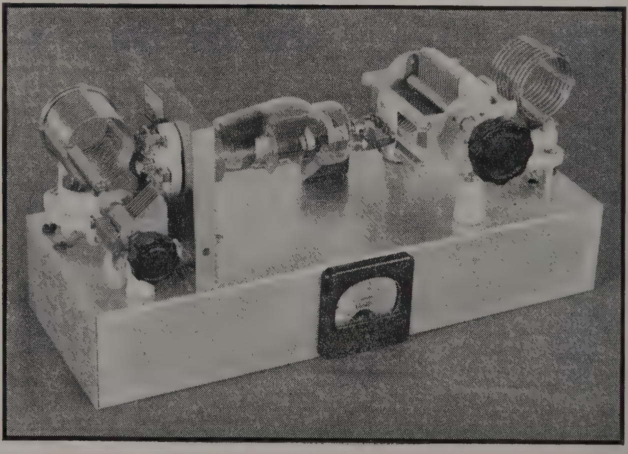
The 10-meter coil consists of 8 turns of no. 16 enamelled, spaced to 1 inch, with a 2-turn link at the ground end.

The 5-meter coil consists of 3 turns of no. 14 enamelled, spaced to 1 inch, with a 1-turn link at the ground end.

All coupling links are wound with no. 18 pushback hookup wire. Link connections are made as shown in Figures 13 and 15.

Operation. When the three stages are properly resonated, only slight readjustment of the tuning knobs will be required to peak up the output when changing bands or moving from one end of the band to the other. When the unit is tuned up initially, care should be taken to make sure the various doublers are working on their second harmonic and not their third harmonic.

Figure 16.
25-WATT UTILITY UNIT, USABLE EITHER AS AN R.F. AMPLIFIER OR OSCILLATOR.
This unit may be used as a crystal oscillator on 160, 80, or 40 meters, or it may be used as an r.f. amplifier (either doubling or straight through) on any band from 10 to 160 meters.



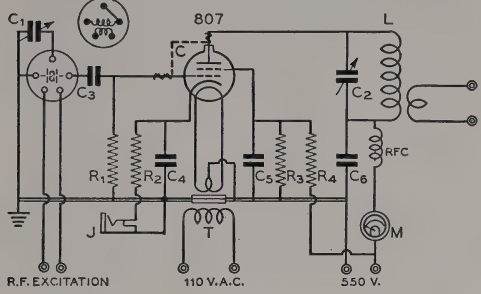


Figure 17.
SCHEMATIC DIAGRAM OF 25-WATT UTILITY UNIT.

- **C_1—100-μμfd. midget condenser**
- **C_2—350-μμfd. condenser, .03″ air gap**
- **C_3—.001-μfd. midget mica**
- **C_4—.01-μfd. tubular paper**
- **C_5—.002-μfd. mica**
- **C_6—.005-μfd. mica, 1000 v.**
- **R_1—100,000 ohms, 1 watt**
- **R_2—500 ohms, 10 watts**
- **R_3—25,000 ohms, 10 watts**
- **R_4—10,000 ohms, 10 watts**
- **L—Manufactured type 50-watt plug-in coils**
- **M—0-100 or 0-150 ma. d.c.**
- **T—6.3 volt 2 amp. fil. trans.**
- **RFC—2.5 mh. choke**
- **J—Closed circuit jack (for key)**

807 UTILITY UNIT

The unit illustrated in Figures 16–18 can be used as a 20-watt crystal oscillator on 160, 80, or 40 meters, or it can be used as a 25-watt straight amplifier on any amateur band from 10 to 160 meters, or it may be used as a 20-watt frequency doubler on any band from 10 to 80 meters.

When used as an amplifier, it may be either link-coupled or capacity-coupled to the exciter tank. When used as a crystal oscillator, the crystal is plugged into the two socket holes indicated in Figure 17. When capacity coupling is used to the unit, the same two connections are used. The coils are provided with a jumper so that when one is plugged into the grid socket the tuning condenser C_1 is cut into the circuit.

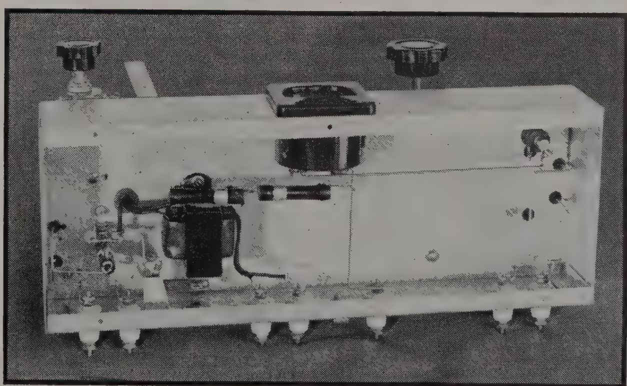
Manufactured type 50-watt end-linked coils are used, the jumper being added. Duplicate coils are not required for grid and plate, as the plate tank makes use of the coil designated by the manufacturer for use on the next higher frequency band (except on 10 meters). This provides a better value of Q in the single ended plate tank circuit, and minimizes the number of coils required to hit several bands. For instance, the "80-meter" coil is used as an 80-meter coil in the grid circuit, but is used as a 160-meter coil when used in the plate circuit. On 10 meters, however, a "10-meter" coil is used in the plate circuit, as the "5-meter" coil requires an excessive amount of capacity to hit 10 meters. Thus, for operation on all bands from 10 to 160 meters, two 10-meter coils would be required, but only one coil for each of the lower frequency bands.

Construction. The unit is built on a chassis measuring 5 inches by 13½ inches by 2½ inches deep. However, a chassis of slightly different dimensions could be used just as well.

Because of the small diameter of the coils, coupling between them is kept to a satisfactory minimum simply by spacing them sufficiently. Electrostatic coupling between grid and plate circuits is prevented by the shield baffle which supports the 807 socket.

The design is such as to accommodate a front panel, the type being left to the preference of the individual constructor. If the constructor is concerned with symmetry of the panel layout, the grid condenser can be raised on higher standoffs so that the shaft height is the same as that of the plate condenser.

Figure 18.
UNDER-CHASSIS VIEW OF UTILITY UNIT.
As will be observed in this illustration, the condensers are mounted so that the shafts project beyond the edge of the chassis, to permit use of a Masonite or metal front panel.



When the unit is used as a crystal oscillator, there is not sufficient feedback to support oscillation except possibly on 40 meters. Hence, when the unit is to be employed as an oscillator, a small piece of insulated wire must be tied to the grid prong of the 807 and run over the shield partition for about 3 inches towards the top of the 807. No more feedback coupling should be used on any band than is required to support stable oscillations, as excessive feedback capacity will result in a high value of r.f. crystal current.

Sufficient cathode bias is provided to limit the plate current to a safe value when no excitation is applied. This permits the unit to be used as an r.f. amplifier after a keyed oscillator or v.f.o.

The unit may be plate-screen modulated with excellent results. The only requirement is that there be sufficient excitation, and that the 807 not be too heavily loaded. The latter applies particularly when modulating the unit as a frequency doubler.

Figure 19.
100-WATT 10-160 METER BAND-SWITCHING EXCITER.
This exciter delivers about 90 watts on 10 meters and well over 100 watts on lower frequency bands.

100-WATT BANDSWITCHING EXCITER OR TRANSMITTER

In Figures 19–23 is shown a unit delivering approximately 100 watts output on all bands from 10 to 160 meters without need for changing coils. Coil switching is incorporated in all three stages of the unit. Using an 814 beam tetrode in the last stage, the output approaches 100 watts on 10 meters and is well in excess of 100 watts on all lower frequency bands.

The oscillator is a conventional 6L6 tetrode type with a tank coil circuit that hits both 80

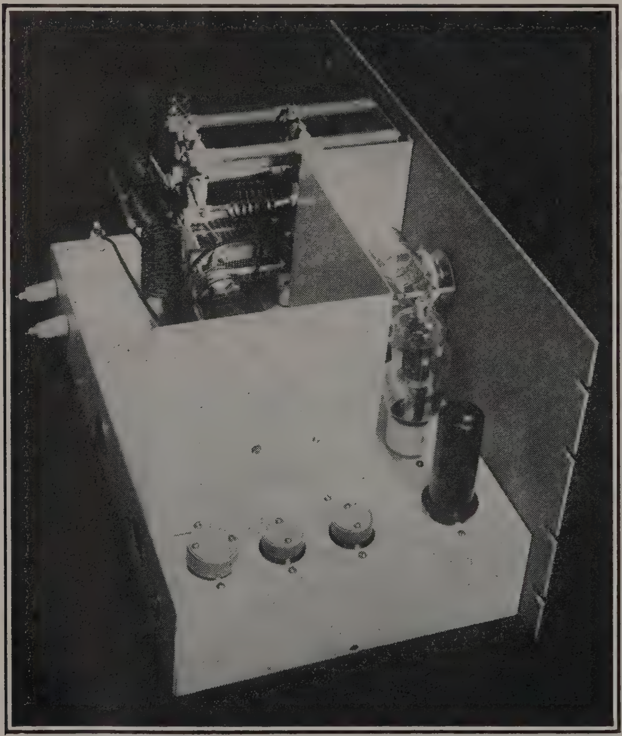


Figure 20.
REAR VIEW OF EXCITER SHOWING OSCILLATOR AND BUFFER.
The 814 plate tank and bandswitch may be seen behind the shield baffle separating the oscillator and buffer from the final stage.

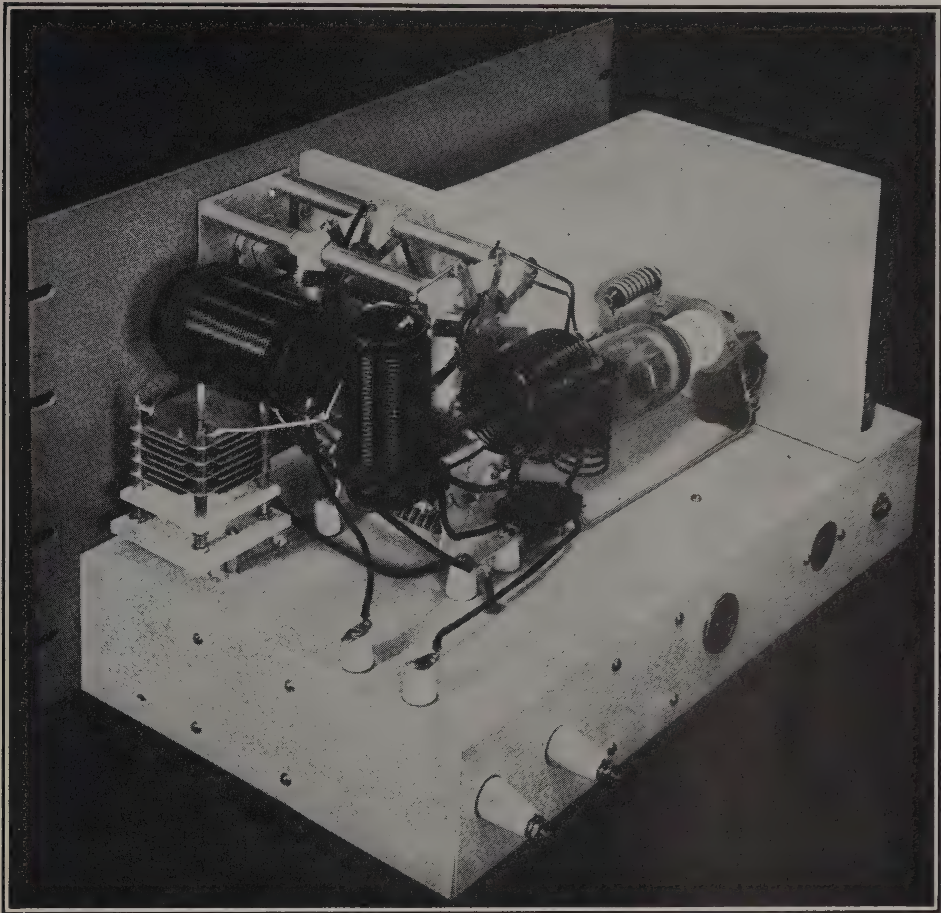


Figure 21.
814 OUTPUT AMPLIFIER OF 100-WATT BANDSWITCHING EXCITER.
Four of the 5 coils are visible in this photo. The 10-meter coil is hidden by the band-switch and tank condenser. The fixed air padding condenser, permanently connected across the 160-meter coil, may be seen in the left foreground.

and 160 meters simply by rotating the condenser plates. To accommodate 40-meter crystals, a shorting switch is connected to a tap on the coil to permit shorting out of sufficient turns to hit 40 meters. The oscillator is run at moderate plate voltage and very low screen voltage to keep the r.f. crystal current low, as not much output is required to drive the 807 stage.

The 807 buffer utilizes a manufactured type midget coil turret to permit output on all bands simply by throwing the coil switch. However, the 10-meter section is not used, inasmuch as the output of the 807 is not sufficient when quadrupling from 40 meters to drive the 814; the latter requires somewhat more excitation on 10 meters than on the other bands, due to relatively high input capacity and resulting tank circuit losses with capacity coupling.

The 814 stage thus may be driven either on 1 or 2 times crystal frequency, and the 814 stage may be run either straight through or as a doubler, the efficiency being nearly as good when doubling as when working straight through as a result of high bias and adequate excitation. Thus, output from the 814 is available on 1, 2, or 4 times crystal frequency.

As the oscillator tank is mounted below the chassis, and the buffer and 814 amplifier tanks

are separated by a shield baffle above the chassis, all three stages are effectively shielded from each other. This results in stable operation when working "straight through."

Three crystal sockets and a crystal switch permit selection of 3 crystals from the front panel. The leads from the crystal sockets to the 6L6 grid via the crystal switch should be made as short and direct as possible. In fact, when a 40-meter crystal is used, it is advisable to place it in the front crystal socket.

The particular method of connecting the low-voltage power supply permits both screen voltage and fixed bias to be obtained for the 814, at the same time providing a desirable compensating action which keeps the 814 grid current from rising to dangerously high values when the load is removed. This compensating effect is obtained with grid leak bias and screen voltage from a series dropping resistor, but is not obtained with ordinary fixed bias and fixed screen voltage. With the system shown, it is important that the B— 500 and B— 1250 volt leads are *not* connected together as is common practice.

The 814 plate tank consists of a husky, 2-gang band switch and 5 coils; data for winding the latter are given in the coil table. To provide a low minimum tuning capacity for good 10-meter efficiency yet sufficient capacity to give a good "Q" on 160 meters, a 100-μμfd. variable condenser is used for tuning, and the 160-meter coil is permanently shunted by a 50-μμfd. air padder.

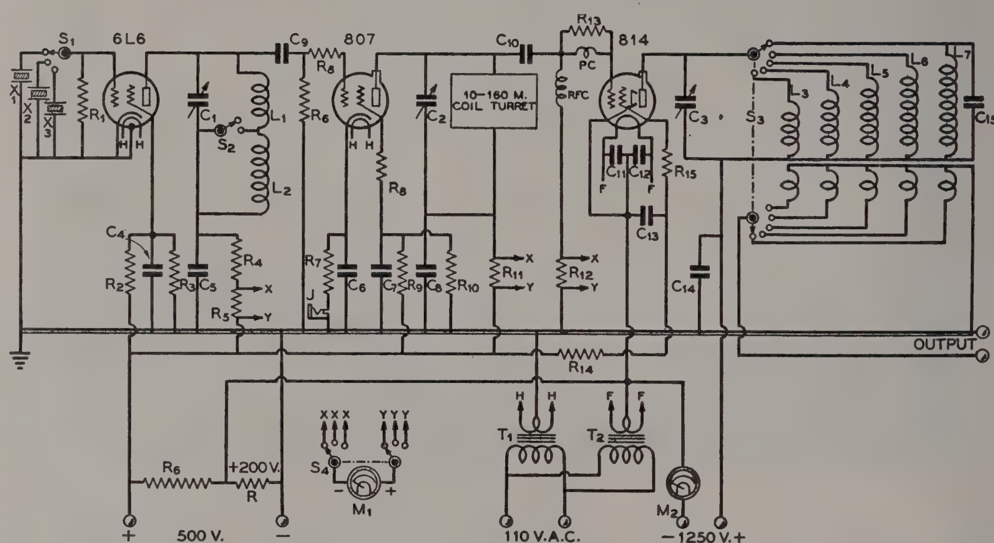


Figure 22.

WIRING DIAGRAM OF BANDSWITCHING 814 EXCITER.

R—4000 ohms, 25 watts
R_1—50,000 ohms, 1 watt
R_2—50,000 ohms, 1 watt
R_3—60,000 ohms, 2 watts
R_4—2500 ohms, 10 watts
R_5—50 ohms, 1 watt
R_6—100,000 ohms, 2 watts
R_7—750 ohms, 10 watts
R_8—50 ohms, 1 watt
R_9—100,000 ohms, 1 watt
R_{10}—50,000 ohms, 2 watts
R_{11}, R_{12}—50 ohms, 1 watt
R_{13}, PC—Parasitic suppressor
R_{14}—5000 ohms, 10 watts
R_{15}—50 ohms, 1 watt
C_1—140-μμfd. midget variable
C_2—100-μμfd. midget variable
C_3—100-μμfd. va-
C_4, C_5, C_6, C_7, C_8—
riable 3000 v. spacing
.003-μfd. midget mica, 1000 v. test
C_9—25-μμfd. midget mica, 1000 v. test
C_{10} — 100-μμfd. mica, 1000 v. test
C_{11}, C_{12}, C_{13}—.003-μfd. midget mica, 1000 v. test
C_{14}—.002-μfd., 5000 v. test
C_{15}—50-μμfd. air padder, 4000 v. spacing
M_1—0-100 ma. d.c.
M_2—0-250 ma. d.c.
T_1—6.3 v. 2 amp.
T_2—10 v. 4 amp.
S_1—Single-pole 3-throw "tone control" switch
S_2—Single-pole 2-throw "tone control" switch
S_3—High power two-gang 5-position bandswitch
S_4—Double-pole rotary meter switch

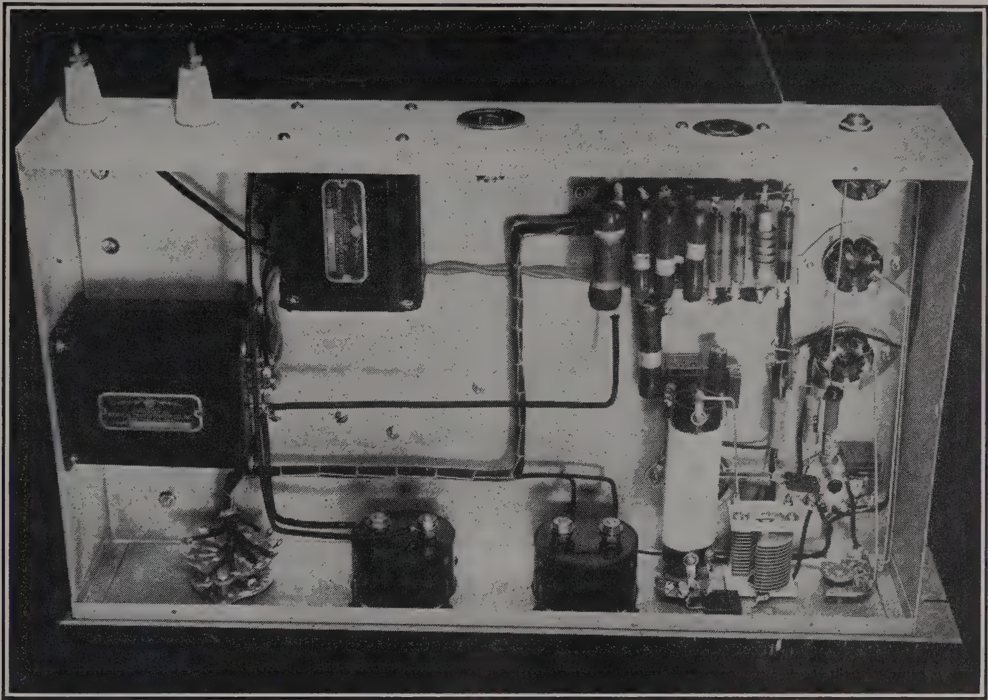


Figure 23.
UNDER-CHASSIS OF THE 100-WATT 814 BANDSWITCHING EXCITER.
The tapped oscillator coil may be seen to the lower right. Most of the resistors are mounted on a terminal strip for the sake of neatness. The meter switch may be seen to the left of the two meters.

COIL DATA
For 100-Watt Bandswitching Exciter

OSCILLATOR COIL
74 turns of no. 22 d.c.c. close-wound on 3½ in. length of 1 in. dia. bakelite tubing, tapped at 24th turn. L_1 is 40-meter section (24 turns).

BUFFER COIL
Manufactured 10-160 meter midget coil turret. 10-meter tap not used.

814 PLATE COILS

10 Meters
6 turns no. 12 enamelled 1⅛ in. dia. spaced to 1⅛ in. (Wound on 1 in. form and form removed.) Link 1 turn at cold end.

20 Meters
11 turns no. 12 enamelled 1⅛ in. dia. spaced to 2 in. (Wound on 1 in. form and form removed.) Link 1 turn at cold end.

40 Meters
18 turns no. 12 enamelled 1¾ in. dia. spaced to 2¼ in. and turns held in place with two celluloid strips cemented to coil with Duco cement. Link 2 turns at cold end.

80 Meters
32 turns no. 12 enamelled close-wound on 3½ in. length of 1½ in. dia. bakelite tubing. Link 3 turns at cold end.

160 Meters
44 turns no. 14 enamelled close-wound on 3¾ in. length of 2 in. dia. bakelite tubing. Link 4 turns at cold end.

All links wound with solid no. 16 having high voltage insulation.

Both to permit shortest possible leads and as a safety precaution, the 814 tuning condenser is set back from the front panel and driven by means of an extension shaft and insulated coupling. The rotor of this condenser is 1250 volts above ground, and the insulated coupling makes it unnecessary to rely upon the insulation of the tuning knob, a bad practice when the voltage is as high as this. The rotor of C_1 and the rotor of C_2 are also above ground, but as the voltage is not particularly high it is only necessary to use knobs with well protected set screws. These two condensers are insulated from the chassis by means of fiber washers.

A 250-ma. meter is permanently connected in the B negative of the 814. Because screen voltage is derived from the low-voltage power supply, the meter reads plate current only; it is not necessary to allow for the screen current when reading this meter. A 100-ma. meter is used to measure current in the various oscillator and buffer circuits by means of a meter switch.

U.h.f. parasitic suppressors are used both in the 807 and 814 stages. These consist of 50-ohm resistors in the 807 control grid and screen grid leads, a 50-ohm resistor in the 814 screen, and a regular parasitic suppressor in the 814 control grid. These suppressors also eliminate all tendency towards instability when working "straight through" on crystal frequency.

The 10- and 20-meter coils must be mounted with very short leads to the 814 coil switch. The leads to the other three coils are not so important. There is bound to be some coupling (both capacitive through the switch and inductive as a result of the unshielded coils) between the high-frequency coils when being used and the low-frequency coils which are left "floating." This is because the 40-meter coil hits fairly close to 10 meters with nothing in shunt, and the 80-meter coil self-resonates near 20 meters. The coils were so designed and placed that the effect is not particularly serious, but small sparks can be drawn from the unused tanks. Fortunately this results in but little loss in efficiency when the 814 stage is loaded, but it does keep the unloaded minimum plate current from being as low as would be the case were plug-in coils used.

On 10 meters the plate current should not be allowed to run over 100 ma. or the plate dissipation will be exceeded. On other bands the plate current should be kept below 110 ma. when doubling, and below 130 ma. when working "straight through." The grid current should be adjusted to about 10 ma. when working straight through, and about 15 or 20 ma. when doubling. Under no conditions should the grid current be allowed to run over 20 ma. when the 814 is loaded.

The grid current to the 814 can be adjusted by detuning the 807, as the 807 is run at such low screen voltage that it will not draw too much plate current or overheat when detuned.

The particular bandswitch used for the 814 plate tank has 6 positions, which leaves one extra. Instead of being left blank, this position is jumpered to the 160 meter switch point. This makes it impossible to remove plate voltage from the 814 by throwing the switch to the unused position. Tubes such as the 814 can be permanently damaged by running them with full screen voltage and no plate voltage.

When mounting the 814 socket, be sure to orient it so that the position of the 814 will correspond to that recommended for horizontal mounting in the manufacturer's application notes. In wiring the 814 socket, be sure to connect the beam forming plates to the *filament return* instead of to ground, as is the more common practice. Connecting the beam forming plates to ground in this case will put 200 volts negative bias on them, greatly reducing the output of the stage.

CHAPTER THIRTEEN

Medium and High Power R.F. Amplifiers

THE amplifiers to be shown in the following pages are typical of those, which, through popular use, have become more or less standard for frequencies up to 30 Mc. and for power outputs of 200 to 800 watts. On frequencies above 30 Mc. special problems arise, and for this reason amplifiers for the higher frequencies are treated separately in Chapters 17 and 19.

Most of the amplifiers illustrated are of the push-pull type, because of the unquestioned superiority of the balanced circuit at high frequencies. The single-ended arrangement finds its widest application in low-power stages, and many such circuits will be found elsewhere in this book. A representative high-power single-ended amplifier is shown later in this chapter, however. It will be noticed that this amplifier is essentially the same as a push-pull amplifier except that one tube and one neutralizing condenser are removed.

Standard Push-Pull Amplifier

Figure 1 shows a standard push-pull amplifier circuit. While certain variations in the method of applying plate and filament voltage and in obtaining bias are sometimes found, the basic circuit remains the same in all amplifiers. All of the push-pull amplifiers illustrated in this chapter use this basic circuit, with such minor variations as are indicated in the descriptions of the individual amplifiers.

Filament Supply. The amplifier filament transformer may be placed right on the amplifier chassis, or it may be located in the power supply, if allowance is made for the voltage drop in the connecting leads. This voltage drop can reach serious proportions where amplifier tubes having low-voltage, high-current filaments are used. In any case, the filament voltage should be the correct value specified by the tube manufacturer when measured *at the tube sockets.* A filament transformer having a tapped primary often will be found useful in adjusting the filament voltage. Where there is a choice between having the filament voltage slightly high or slightly low, the higher voltage is preferable. If the amplifier is to be greatly overloaded, a filament voltage slightly higher than the rated value will give greater tube life.

Plate Feed. The series plate-voltage feed shown in Figure 1 is the most satisfactory method for push-pull stages. This method of feed puts high voltage on the plate tank coil, of course, but since the r.f. voltage on the coil is in itself sufficient reason for protecting the coil from accidental bodily contact, no additional protective arrangements are made necessary by the use of series feed. On the low frequency bands a plate r.f. choke will not always be required with this type of amplifier. However, one is usually desirable on the higher frequency bands, and as the choke does no harm in any case its incorporation is recommended.

The insulation in the plate-supply circuit should be adequate for the voltages encountered. In c.w. and grid-modulated stages, the insulation of the r.f. choke and wiring should be capable of withstanding voltages at least as high as the plate voltage. Where plate modulation is used, the insulation should be able to withstand at least twice the d.c. plate voltage. If the plate-current meter is placed in the positive lead, it, too, must have adequate insulation between the movement and case.

Grid Bias. The recommended method of obtaining bias for c.w. or plate modulated telephony is to use just sufficient fixed bias to protect the tubes in the event of excitation failure and obtain the rest from a grid leak. However, the grid leak may be returned directly to the filament circuit if an overload relay is incorporated in the plate circuit, the relay being adjusted to trip immediately when excitation is removed. For grid modulation it is necessary that all the bias be obtained from a fixed source; this makes a grid leak impracticable for this class of service.

The grid leak R_1 serves effectively as an r.f. choke in the grid circuit because the r.f. voltage impressed upon it is very low, and no grid r.f. choke is required when a grid leak is used. However, if no grid leak is incorporated, as would be the case for fixed bias for grid modulation, an r.f. choke should be substituted for R_1. This choke should have a different value of inductance than the choke in the plate circuit, since equal values of inductance will often cause a low-frequency parasitic oscillation. Should low-frequency parasitics persist with different sized chokes, a 200-ohm, 10-watt wirewound resistor may be placed in series with the grid choke. The resistor will suppress the oscillation without otherwise affecting the operation of the amplifier.

Metering. It will be noticed in Figure 1 that M_2 is placed in the negative-to-filament return rather than in positive high-voltage lead. This is a safety precaution. When connected as shown in the diagram, M_2 will read plate current only, as M_1 is returned to the "hot" side of M_2 instead of to the negative plate lead. This will require an extra external lead if fixed bias is used, as the positive of the bias supply cannot be connected to the negative plate voltage under these conditions without resulting in a short across M_2.

When measuring current in the filament return of filament type tubes, it is necessary that the stage have *either* an individual power supply or else a filament supply which is not used to supply any other *filament type* tubes (heater tubes may be operated from the same filament supply). If this requirement is not met, a meter jack will read the current being drawn by more than one stage at the same time. If desired, meter jacks or a switch may be substituted for the individual meters in Figure 1.

Plate Circuit. In the circuit shown in Figure 1, the rotor of the plate tank condenser is left "floating" (ungrounded). This permits a tank condenser of less spacing to be used, as there is no d.c. impressed across it. When the rotor is "floating" it is imperative that the amplifier be symmetrical from a physical standpoint, and that the coupling to the external load be symmetrical. Because the rotor will be at high d.c. potential if the condenser *should* arc over, it is advisable to use an insulated coupling between the rotor shaft and the tuning dial or knob.

In cases where it is impossible to obtain equal loading of the two tubes in the push-pull amplifier, it may become necessary to ground the rotor of the plate condenser through a by-pass condenser. If the stage is plate modulated, it may then be necessary to connect the rotor of the condenser to the modulated plate voltage lead directly or through a 25,000- to 100,000-ohm resistor to allow the rotor to follow the modulation voltage. The resistor may be a 1-watt carbon unit. When the resistor is used, the by-pass condenser between rotor and ground should not have a capacity of greater than .001 μfd. A larger condenser causes excessive phase shift between the modulated voltage and the voltage on the rotor at the higher audio frequencies, and thus increases the instantaneous voltage between rotor and stator. There is no restriction on the size of by-pass condenser when a direct connection is used between the rotor and modulated voltage, except that the condenser must not be so large that it by-passes an appreciable portion of the modulation.

Because of the high minimum capacity of tuning condensers having sufficient maximum capacity for proper 160-meter operation, it is good practice to use a split stator plate tuning condenser just sufficiently large for 40-meter c.w. operation (about 75 $\mu\mu$fd. per section for commonly used ratios of plate voltage to plate current) and then use external plug-in fixed padding condensers for 80- and 160-meter operation. The cost is about the same as for a split stator condenser having sufficient capacity for 160-meter phone operation, and the efficiency on 10 and 20 meters is higher because of the lesser bulk and minimum capacity of the tuning condenser. In the low and medium power range, fixed air padders are the least expensive; for high power operation, fixed vacuum condensers are about as economical as the regular air types. Recommended values of tank circuit capacity for different bands and applications are given in Chapter 7.

For high-power operation on 10 and 20 meters, a fixed capacitance is sometimes used in conjunction with a variable inductance to replace the more common type of plate tank consisting of a fixed inductance and variable capacitance. This is permissible in the circuit of Figure 1 so long as the fixed tank condenser is symmetrically constructed. It is not advisable to substitute a single-section variable condenser of twice the spacing and half the per-section capacity for C_2 because it would upset the symmetry of the circuit; the rotor (frame) consists of so much more metal than the stator that there would be considerable unbalance with this type of condenser.

Plate tank coils for medium- and high-power amplifiers may be wound of bare or enameled copper wire (no. 14 or larger) or of the smaller sizes of copper tubing. Coils for 28 Mc., and sometimes for 14 Mc., may be made self supporting when wound with the larger sizes of

wire or with copper tubing. For lower frequencies, high-grade ceramic forms may be used, or the coils may be made mechanically rigid by cementing the turns to celluloid strips.

Grid Circuit. As the power in the grid circuit is so much lower than in the plate circuit, it is customary to use a split stator grid condenser with sufficient capacity for operation on the lowest frequency band, and also to ground the rotor. A physically small condenser has a greater ratio of maximum to minimum capacity, and it is possible to get a grid condenser that will be satisfactory on all bands from 10 to 160 meters without need for external auxiliary capacitors. As both r.f. and d.c. voltages are relatively low in the grid circuit, the rotor of the condenser can be grounded without increasing the cost appreciably, as very little more spacing will be required and the condenser is relatively small anyhow (in comparison with the plate tank condenser). Grounding of the rotor simplifies mounting of the condenser, and also provides circuit balance and insures electrical symmetry. It also retards u.h.f. parasitics by by-passing them to the ground in the grid circuit.

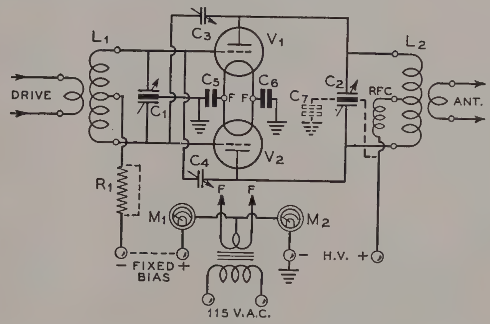
Coils for the grid circuit may in most cases be mounted on small jack-bar or tube-base type supports. Wire sizes up to no. 14 will be suitable for driving powers up to 100 watts. To restrict the field and thus aid in neutralizing, the grid coils should be physically no larger than absolutely necessary.

Layout. The most important consideration in constructing a push-pull amplifier is to maintain electrical symmetry on both sides of the circuit. Of utmost importance in maintaining electrical balance is the stray capacity between each side of the circuit and ground.

Large masses of metal placed near one side of the grid or plate circuits can cause serious unbalance, especially at the higher frequencies, where the tank capacity between one side of the tuned circuit and ground is often quite small in itself. Capacity unbalance most often occurs when a large plate or grid coil is located with one of its ends close to a metal panel. The solution to this difficulty is to mount the coil parallel to the panel to make the capacities equal from each end to ground, or to place a large piece of metal opposite the "free" end of the coil to accomplish the same purpose.

Wherever possible, the grid and plate coils should be mounted at right angles to each other. If this is not practical, the coils should be separated as far as possible. A small

FIGURE 1. STANDARD PUSH-PULL R.F. AMPLIFIER CIRCUIT.



The mechanical design must be symmetrical and the output coupling must be evenly balanced. Individual meters may be substituted for the two meter jacks. If a grid r.f. choke is substituted for R_1 for fixed bias operation, a 200-ohm wire-wound resistor should be placed in series if a low frequency parasitic oscillation occurs.

C_1—Approx. 1 $\mu\mu$fd. per section per meter of wavelength. 1000 volt spacing for HK-54, 35T, T55, 812, 808, etc. 2000-volt spacing for 100TH, HK-254, HF-200, T200, etc.

C_2—Refer to tank condenser data and Q charts in Chapter 7 for capacity and spacing.

C_3, C_4—Suitable neutralizing condensers, 50% greater air gap than C_2. Maximum usable capacity should be slightly greater than grid-plate capacity of tubes.

C_5, C_6—.002 μfd. or larger.

C_7—Not over .004 μfd.

R_1—Of such value that normal grid current for tubes will produce enough voltage drop to make a total of twice cut-off bias including any fixed bias. Higher resistance can be used with slight increase in efficiency if reserve of excitation is available. Wattage rating equal to I^2R. The resistor may be omitted entirely and fixed bias equal to twice cut-off or more may be used, if desired.

RFC—2.5-mh. r.f. choke designed for all-band operation, of suitable d.c. rating. Not always found necessary.

T_1—Filament transformer of suitable voltage and current rating. Tapped primary desirable, especially if transformer is located some distance from the amplifier.

M_1, M_2—Suitable grid- and plate-current meters.

amount of coupling between the two coils is not in itself greatly detrimental, since it can usually be balanced out by the neutralizing circuit, but the coupling will vary when coils are changed and it will be necessary to readjust the neutralizing when changing bands.

All r.f. leads should be made as short and direct as possible, of course. The leads from the tube grids and plates should be connected directly to their respective tank condensers, rather than to the coils. The connections between the coils and condensers should be of wire or tubing at least as large as that used in the coils themselves. Plate and grid leads to the coils need not be as heavy as the tank circuit leads, but the use of flexible tinned "braid" is to be avoided wherever possible, since the braid has rather high r.f. resistance. Where braid must be used to provide flexibility when connecting to grid or plate caps on the tube envelopes, a short piece of the braid should be soldered to the end of a solid-wire lead and the wire used for the major portion of the connection. A still better method is to use thin, flexible copper strip for leads to the tube caps.

Many of the troubles so often associated with neutralizing can be obviated by running the neutralizing leads directly to the tube grids and plates entirely separately from the grid and plate leads to the tank circuits. Having a portion of the plate or grid connections to their tank circuits serve as part of a neutralizing lead, or vice versa, can often result in apparently mysterious neutralizing troubles. The importance of eliminating the common leads is shown by the fact that certain tubes designed for u.h.f work have entirely separate leads brought out from the elements for tank-circuit and neutralizing connections.

Excitation. The excitation requirements for high- and medium-power amplifiers vary so widely that it is difficult to make definite general statements of the driving power which should be provided. However, a good average figure for the excitation power to modern triodes in a class C amplifier is that it should approximate 10 per cent of the expected power output of the stage. Where extremely high efficiency in the amplifier is desired, the excitation may have to be as high as 20 or 30 per cent of the power output, and where the amplifier can be operated class B for c.w. purposes, the excitation power may sometimes be as low as 5 per cent of the power output. Pentodes and tetrodes generally require less excitation than triodes of equivalent plate dissipation, but their high input capacities make them more difficult to excite, and the required driver power output, while usually considerably less than that required for an equivalent triode, is not always as low as is sometimes thought. Excessive excitation to pentodes or tetrodes will often result in reduced power output and efficiency, however. Except in the case of pentodes and tetrodes, it is best to err on the side of excessive excitation, since the surplus will do no harm and a scarcity of excitation will cause a loss in output and efficiency.

The best rule to follow in adjusting the excitation is to use all the excitation available, and then adjust the bias until the grid current is at the rated operating figure given by the tube manufacturer. In push-pull or parallel stages, the current should be twice the value given for one tube, of course. If a fixed bias supply is used, and the grid current is excessive with the bias voltage set at its maximum value, additional grid-leak bias should be introduced to reduce the current to its rated value. The actual bias will then be equal to the fixed supply voltage plus the voltage contributed by the IR drop in the bias resistor. Where grid-leak bias alone is used, the bias will simply be equal to the IR drop in the resistor (grid current in amperes times grid-leak resistance in ohms). A combination of grid-leak or fixed bias and cathode-resistor bias will give a total bias equal to the sum of the voltage contributed by the IR drop in the cathode resistor and the voltage supplied by the grid-leak or bias supply. When computing the drop across a cathode resistor it must be remembered that not only the plate current, but also the grid current and the screen current, if any, flow through the cathode resistor.

The above general rule for adjusting the bias to conform with the excitation will sometimes lead to a value of bias that is too low for class C 'phone operation, when the excitation is low. In such cases it will not be possible to get more bias by raising the value of the grid leak, since the grid current drops as the resistance is increased, and little increase in bias is obtained.

Single-Ended Stages. Most of the preceding discussion, except the section on circuit balance, applies equally well to single-ended as well as push-pull stages. Even in single-ended stages, however, it is desirable to maintain capacity balance to ground from both sides of the plate circuit when a split-stator plate condenser is used to obtain neutralizing voltage. In the single-ended stage, capacity balance is obtained by *adding* a capacity from the "free" end of the plate tank to ground, to make up for the tube's plate-filament capacity across the other side of the circuit. The balancing capacity may be obtained by placing an actual condenser equal to the plate-filament capacity between the free end of the tank circuit and ground, or in the case of tubes having a low plate-filament capacity, by locating the plate

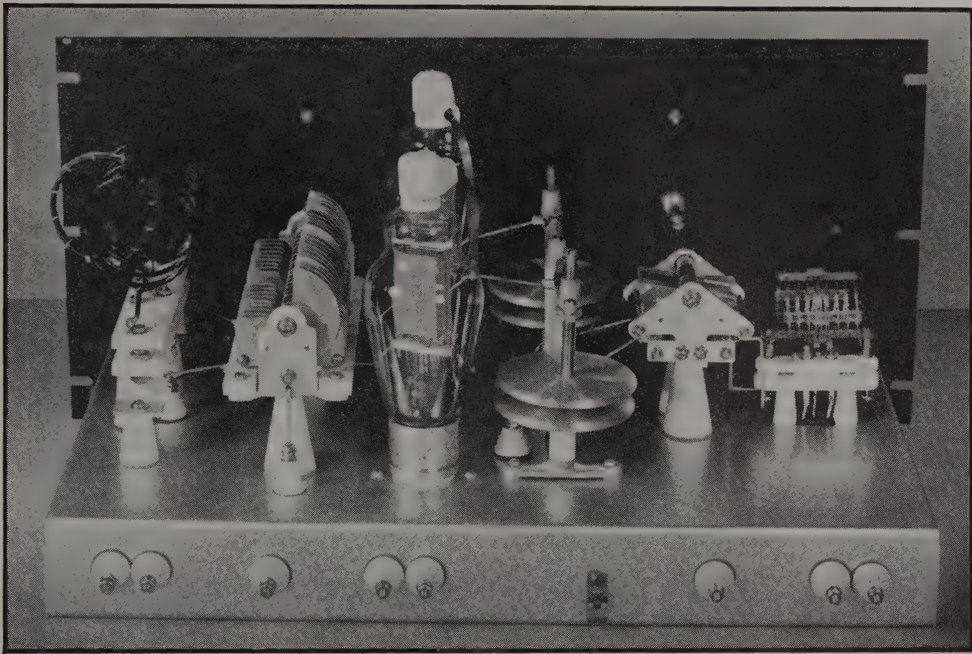


Figure 2.
PUSH-PULL 8005 AMPLIFIER.
This amplifier will deliver an output of 440 watts on c.w., or 340 watts when plate modulated. The amplifier is designed for mounting in a standard rack, and its layout is perfectly straightforward. Note that the grid and plate coils are mounted at right angles to each other to reduce coupling between the input and output circuits.

coil so that its free end is close to the chassis or panel. An example of the latter system is shown later in this chapter.

Because the single-ended circuit is not inherently balanced to ground, the necessity of obtaining proper ground connections is all-important. The filament, grid, and plate by-pass condensers should all be returned by the shortest possible *separate* leads to a common point on the chassis. Grounding these condensers to widely separated points on the chassis, or to a common ground bus, is quite likely to lead to difficulties with feedback or instability due to coupling between the various circuits in the chassis or common lead. The connection between the filament by-pass condensers and chassis should be as short as possible, with the other by-pass condensers grounded where the filament by-passes are connected to the chassis. At the higher frequencies, it may even be advisable, in some cases, to connect the grid and plate by-pass condensers right to one of the filament terminals on the tube socket, and then by-pass that side of the filament to the chassis with the shortest possible leads and also by-pass the two sides of the filament together at the socket.

The pictorial illustrations in this chapter will be found useful for the purpose of furnishing ideas for possible mechanical layouts. All of the arrangements shown permit very short r.f. leads, but it is not necessary to use the particular tubes specified in each case for the particular physical layout illustrated. For instance, with very slight modifications in the amplifier, 35T's, HK54's, HY-51's, 812's, or T55's could be used in the amplifier pictured in Figures 2 and 3 by providing the proper grid leak and filament transformer. Much of the enjoyment to be obtained from amateur radio comes from experimenting with the design of amplifiers such as these, and the units in this chapter are shown simply as samples of arrangements which have given good results. The individual constructor will often find it advisable and instructive to alter the designs to suit components which he has on hand, or to incorporate different tubes than those shown.

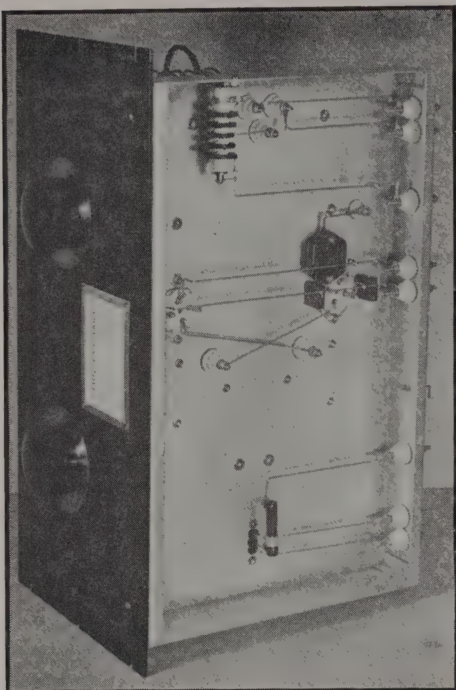


Figure 3.

UNDER-CHASSIS VIEW OF THE 8005 AMPLIFIER.

The filament and plate by-pass condensers, plate r.f. choke, and the grid resistor are mounted under the chassis. The crossed grid leads from the above-chassis neutralizing and grid circuit may be seen in this photo.

Push-Pull 8005 Amplifier

An amplifier using push-pull 8005 tubes is shown in Figures 2 and 3. This amplifier is capable of giving a power output of 440 watts for c.w. service, or 340 watts when plate modulated (ICAS ratings).

The amplifier is constructed on a 17 x 10 x 2-inch chassis, which is surmounted by a standard relay-rack panel 8¾ inches high. From left to right across the chassis are located the grid coil, grid condenser, neutralizing condensers, tubes, plate condenser, and the plate coil. To shorten the length of the grid and plate leads, the two tank condensers are raised above the chassis on 1¾-inch standoff insulators. The grid coil jack bar is raised above the chassis a small amount by mounting it on the insulators supplied by its manufacturer. The grid coils are "50-watt" manufactured units. In the plate circuit, a standard manufactured jack-bar assembly is used, with home-wound coils mounted on a matching plug assembly. The jack bar is supplied with mounting feet, and these are supported 1 inch above the chassis by means of standoff insulators. Ten- and 20-meter coils for the plate circuit may be wound "on air" and supported from the plug-in base. The 40-meter coil should be wound on the 2½-inch diameter ceramic form manufactured for use with the plug bar.

Terminals at the rear of the chassis are provided for connection of the various supply voltages, the link from the exciter, and connections to the antenna. The d.c. portion of the plate circuit is wired as shown by the dotted lines in Figure 1.

For c.w. use, the plate voltage may be as high as 1500 volts, and the plate current as high as 400 ma. The grid current should be 65 ma. for the two tubes, while bias may be obtained from a fixed supply of 130 volts or from a resistor of 2000 ohms.

When plate modulated, the plate voltage should be 1250 volts for maximum output, while the plate current may be as high as 380 ma. A fixed supply of 195 volts or a 3500-ohm resistor may be used for grid bias. The grid current should be 60 ma. An exciter having an output of 30 to 40 watts will be adequate for exciting the amplifier on either 'phone or c.w.

35-TG Amplifier

The amplifier shown in Figure 4 follows the general diagram of Figure 1, as do all the push-pull amplifiers in this chapter. Since the grid terminals on the 35-TG's are on the side of the glass envelope, it is convenient in an amplifier using these tubes to place all the r.f. components and r.f. wiring above the chassis. To help keep the length of the plate leads to a minimum, the tube sockets are mounted about ½ inch below the chassis. Half-inch sleeves over the socket mounting screws serve to hold the sockets firmly in position. The plate connections to the tank condenser and neutralizing condensers are made by means of thin, flat copper strip, to provide the necessary flexibility. Radiator-type connectors are used on the tube plate terminals.

The neutralizing condensers are a type ordinarily intended to mount with the plane of their plates vertical. However, for use in this amplifier, it is more convenient from a wiring standpoint to have them mount at right angles to their usual position. This is done by screwing what should be the bottoms of the two condensers together, with a strip of scrap chassis metal placed between the two bottoms. The strip of metal is allowed to extend past the bottom on one side, and the extension bent

over to form a mounting foot for both condensers.

Spacing sleeves 1 inch long are used to support the grid-coil socket, which is an ordinary 5-prong Isolantite socket. The grid coils are "50-watt" manufactured units. The plate coils are also of the manufactured type; they have a power rating of 500 watts. A swinging antenna-coupling coil, which is part of the plate coil jack-bar assembly, allows the antenna loading to be adjusted to the proper value.

Standoff insulators are provided at the rear of the chassis for connections to the plate, filament, and bias supplies, and for link leads from the exciter. For c.w. use, the plate voltage may be as high as 2000 volts, and the plate current run up to 250 ma. without difficulty. The plates of the 35-TG's will run red at normal plate dissipation. When the amplifier is plate modulated, the plate voltage may be as high as 1500 volts, and the plate circuit loaded to 250 ma. When plate modulation is used, the rotor of the plate tank condenser should be connected to the modulated high voltage and by-passed to ground, as shown by the dotted lines in Figure 1.

The grid current for the two tubes should be 60 ma. under normal operating conditions. Bias may be supplied by either a resistor or a fixed supply, or a combination of each. For c.w. operation at 2000 plate volts, the fixed bias should be at least 175 volts, or a 3000-ohm resistor may be used with 60 ma. of grid current flowing. In plate-modulated service, the bias may be provided by a fixed supply of at least 120 volts, or by a resistor of at least 2000 ohms, with 60 ma. of grid current. The exciter should have an output of 40 to 50 watts.

HK-254 Amplifier

An amplifier capable of handling the full legal amateur input of 1 kw. in c.w. use is shown in Figure 5. High-efficiency operation of the HK-254 tubes is necessary if they are to be kept within their plate dissipation rating

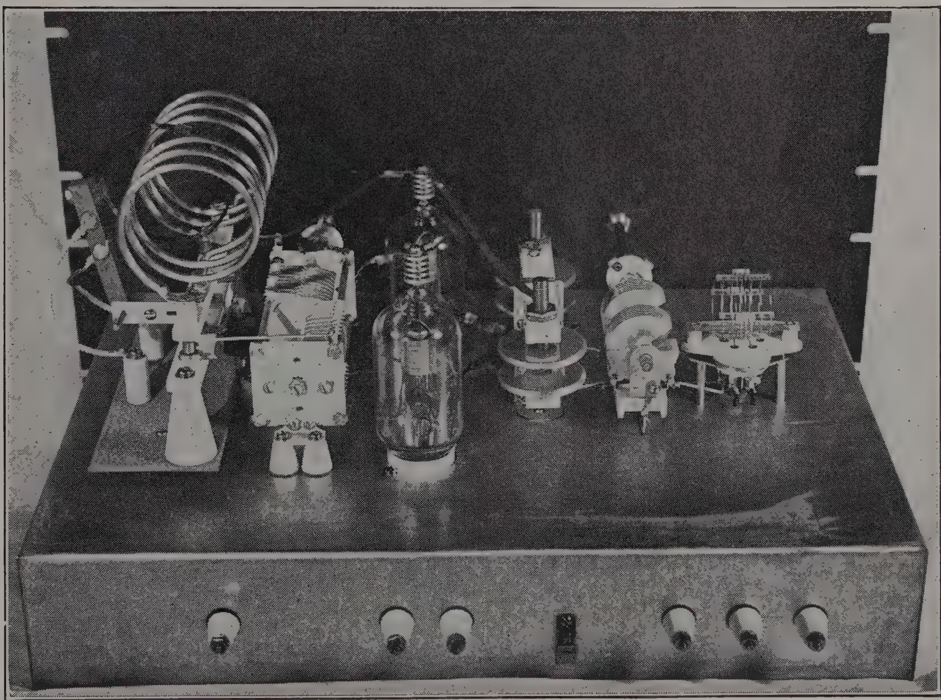


Figure 4.
400-WATT 35-TG AMPLIFIER.
This amplifier is capable of supplying a 400-watt signal on c.w., and nearly that much on 'phone. Any exciter having an output of 40 to 50 watts will serve to drive it. The terminals at the rear of the chassis are for plate, bias, and filament voltage connections. The antenna or antenna coupler is connected to the variable link terminals on the plate coil assembly.

at 1 kw. input, and the amplifier has been constructed with this requirement in mind. The arrangement of parts shown in the photo allows the average length of the r.f. leads to be kept down to about 1 inch. None of the r.f. leads is over $2\frac{1}{2}$ inches long.

The amplifier is constructed on a 17 x 12 x 3-inch chassis, most of the wiring being above the chassis. The arrangement of parts proceeds in an ordinary manner from left to right along the chassis. Near the left end of the chassis is a 5-prong socket for the plug-in grid coils, which are wound on standard $1\frac{1}{2}$-inch receiving-type forms. To the right of the coil socket is the split-stator grid condenser, which is surmounted by the neutralizing condensers. Next in line are the tubes, which are mounted near the front and back of the chassis, with their sockets below the chassis. The plate tank condenser is located directly to the right of the tubes, and it is mounted upside down to bring its stator terminals near the tube plate connections. Large jack-top standoff insulators ($4\frac{1}{2}$ inches) are used to support the plate coil, and these bring the coil terminals up next to the condenser stator terminals. Through-panel standoff insulators in the rear drop of the chassis provide terminals for the connection of plate, grid, and filament voltage, and the link from the exciter.

For c.w. use, the plate voltage should be 2500 volts, and the plate current may be as high as 400 ma. The grid current should be 80 ma. for both tubes. Bias may be from a fixed source of 250 volts or more, or from a resistor of at least 3000 ohms, with 80 ma. of grid current flowing. In plate-modulated service, the plate voltage, grid bias, and grid current requirements remain as above, but the plate current should be held below a maximum of 350 ma. An exciter power output of 75 watts will be adequate for 'phone or c.w. service.

TW-150 Amplifier

The amplifier pictured in Figure 6 is suitable for operation on 5-, 10-, or 20-meter 'phone or c.w., and 40-meter c.w., with an input of 1 kw. The amplifier is intended to mount behind a standard relay-rack panel, as are all the other amplifiers shown in this chapter. However, the amplifier is shown in Figure 6 with the panel removed, to help in showing the method of construction. The panel goes across the right side of the amplifier in the photo, with the shafts from the two tank condensers projecting through the panel. To avoid confusion in the following discussion, the location of the parts will be described as though the side facing the observer were the front; actually, however, it is the left side of the amplifier.

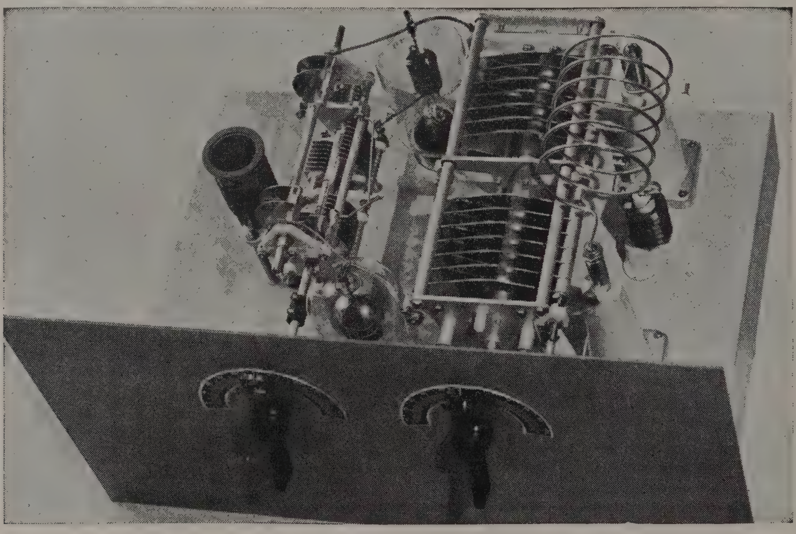


Figure 5.

800-WATT AMPLIFIER WITH HK-254'S.

This high powered amplifier will take a full kilowatt input on c.w. if 75 watts of excitation is available. High efficiency operation is required, as otherwise the plate dissipation rating of the tubes will be exceeded at 1 kw. input. The HK-254's are fed 2500 volts and loaded to 400 ma. The physical layout illustrated permits an average r.f. lead length of slightly over 1 inch; no lead is over 2½ inches.

An 8 x 17 x 3-inch chassis is used for the amplifier, and the tubes are located toward the front, protruding through 3-inch diameter holes. The sockets are mounted on the chassis bottom plate. Immediately behind the TW-150's, and elevated so that stator tie rods are about an inch below the tube plate terminals, is the plate tuning condenser. Between the tubes, and forward from the condenser, are the neutralizing condensers, positioned so that short, direct leads are featured between lower plate connecting lugs and tube grid terminals.

A length of Mycalex strip between and above the neutralizing condensers and fastened to them supports the plate circuit r.f. choke, and two shorter lengths similarly fastened and protruding to right and left over the tube grid caps support the 3-inch ceramic pillars which

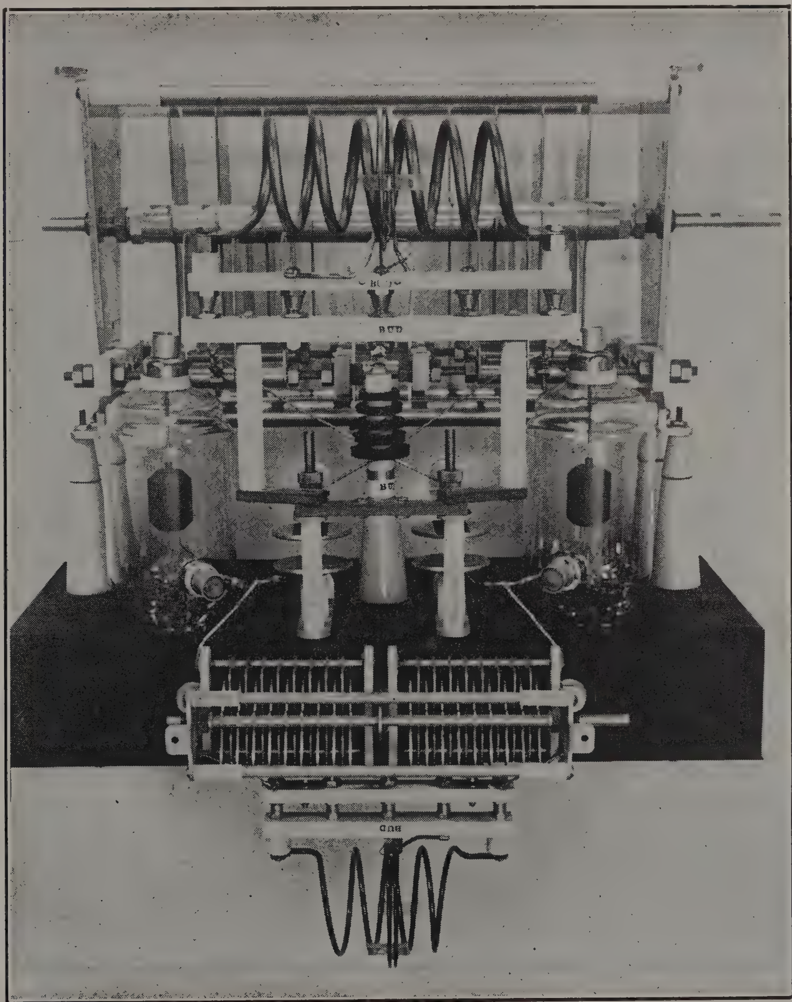


Figure 6.
800-WATT AMPLIFIER OF UNUSUAL DESIGN.
This "skeletonized" push-pull TW-150 amplifier is capable of handling an input of 1 kilowatt, with a plate efficiency of between 75 and 80 per cent. The amplifier is intended to be mounted behind a 19-inch rack panel with the tuning controls being connected to the grid and plate condensers by means of insulated couplings. Note how the r.f. leads are kept to extremely short lengths, for an amplifier of this size, through the use of the unconventional chassis arrangement.

in turn support the plate tank mounting assembly. The grid tuning condenser mounts on the chassis front drop, while the grid coil and the grid leak mount below chassis on the bottom plate.

Connections from the plate coil and the tube plate cap terminals to the tank condenser run directly to the stator tie rod. The tie at the frame is made with clips which come with adjustable resistors of the 50-watt type. Crossover neutralizer connections are made to the inside ends of these tie rods. Grid connections from neutralizers to tubes, tubes to tuning condenser, and condenser to coil assembly are all unusually short for kilowatt construction.

The high-voltage plate supply should provide a maximum of 3000 volts at 330 ma. for 990 watts input, 'phone or c.w. The grid current should be 80 ma. under operating conditions, and bias may be supplied by a fixed source of 260 volts, or from a resistor of approximately 3500 ohms. About 75 watts of excitation will be needed by the amplifier at the higher frequencies.

Single-Ended 152-T Amplifier

Although it is usually preferable, both from a standpoint of efficiency and tube cost, to use a push-pull amplifier for medium and high

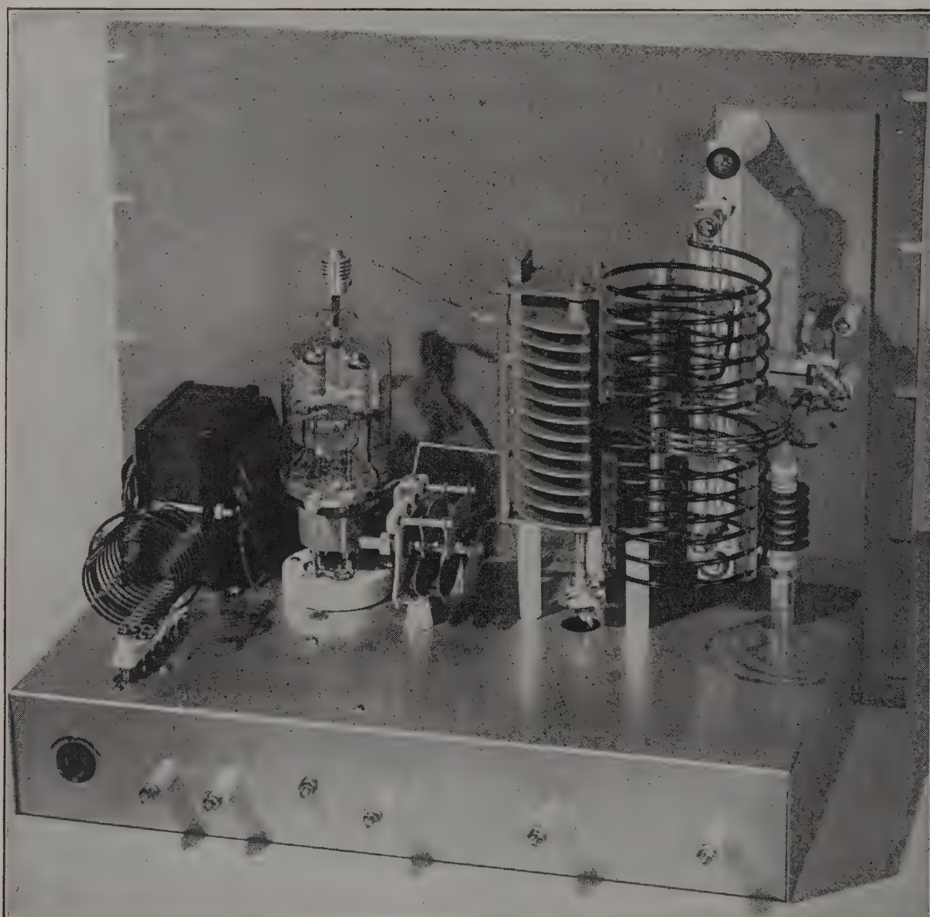


Figure 7.
750-WATT SINGLE-ENDED 152-T AMPLIFIER.
This amplifier will give an output of 750 watts on c.w. when run at an input of 900 watts. Mounting the plate tank circuit vertically, with the neutralizing condenser between the tube and tank circuit, aids in preserving circuit balance in the single-ended plate neutralized stage. The filament transformer is located on the chassis.

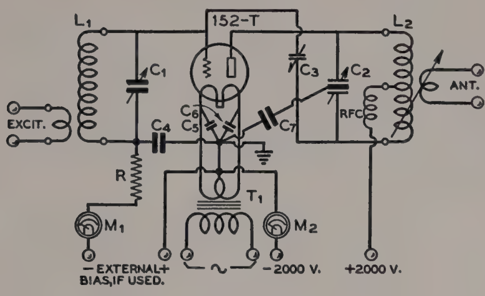


Figure 8.

WIRING DIAGRAM OF TYPICAL SINGLE-ENDED AMPLIFIER.

This diagram is for the amplifier illustrated in Figures 7 and 9. The tank condenser illustrated is for 10, 20, and 40 meter operation only. The 152-T has two separate sets of filament connections brought out to the socket. These may be connected in series and operated from a 10-volt transformer, as shown in the diagram, or they may be connected in parallel and supplied from a 5-volt transformer.

- C_1—100-$\mu\mu$fd., .070″ spacing
- C_2—50-$\mu\mu$fd. per section, .171″ spacing
- C_3—9-$\mu\mu$fd., .238″ spacing
- C_4, C_5, C_6—.002-μfd. 1000-volt mica
- C_7—.002-μfd. 5000-volt mica
- R—7000 ohms, 50 watts. If fixed cut-off bias is used (250 v.) this resistor may be reduced to 3000 ohms, 20 watts
- L_1—"100-Watt" manufactured coil
- L_2—"1 Kw." manufactured coil with swinging link mounting
- RFC—2.5 mhy., 500 ma.
- T_1—10 volts, 7 amp. (if filaments are connected in parallel T_1 may be 5 volts, 14 amp.)
- M_1—0-100 ma.
- M_2—0-500 ma.

power transmitters, some circumstances, such as the availability of a large tube or desire to couple an extremely unbalanced load to the amplifier, often make it advisable to use a single-ended output stage.

The 152-T amplifier illustrated in Figures 7 and 9 and diagrammed in Figure 8 is typical of single-ended amplifier circuits. The circuit shown is also applicable to other tubes having a relatively low output capacity. Where the tube's output capacity is high, however, spe-

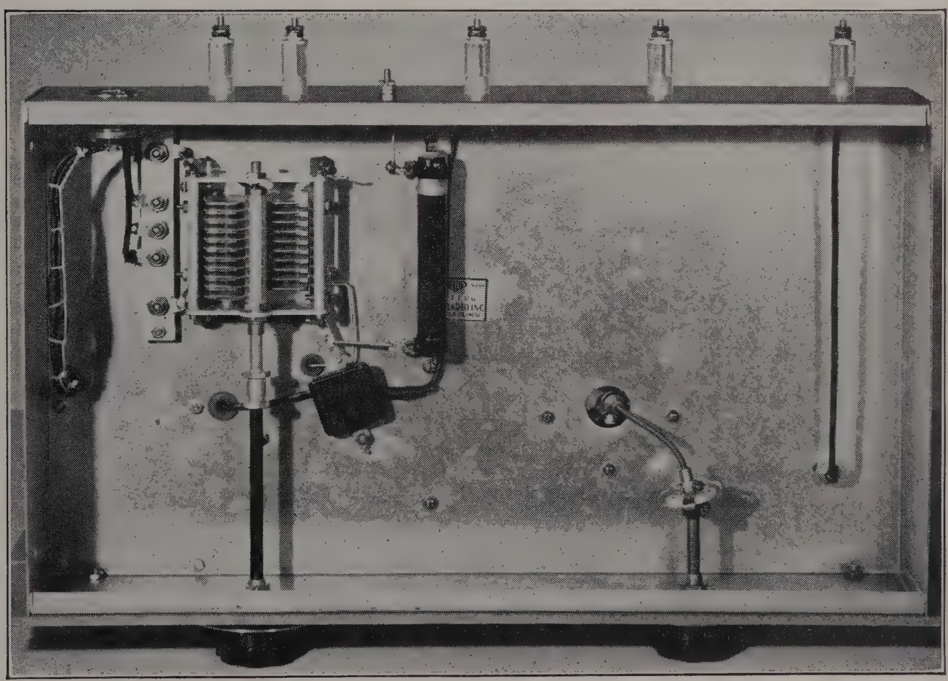


Figure 9.

UNDER-CHASSIS VIEW OF THE SINGLE-ENDED AMPLIFIER.

The grid condenser and bias resistor are located under the amplifier chassis. Note the flexible shaft which allows the vertically mounted plate condenser to be controlled from the panel.

cial circuit arrangements are necessary to assure correct neutralization, as explained in Chapter 7.

For a single-ended amplifier, it is necessary that the rotor of the plate tank condenser have an r.f. return to ground. This may be done either by grounding the rotor of the condenser directly, or grounding it through a by-pass condenser. If it is grounded directly, the tank condenser must have somewhat more spacing because both d.c. and r.f. are impressed across it. If it is by-passed to ground through a high-voltage mica condenser of .002-μfd. to .004-μfd. capacity, as shown in the diagram, there will no longer be impressed, under steady carrier conditions, anything between the condenser plates except r.f. voltage. However, transient peaks will be impressed across the variable condenser during plate modulation or primary keying of the stage. Hence, to remove all voltage from the variable condenser except the r.f. voltage across the coil, the rotor of the condenser should be connected directly or through a resistance to the positive high voltage lead, as previously described in this chapter. There can then be no a.f. or d.c. transient voltage impressed across the condenser sections, because both the rotor and the stator sections are at the same potential except with respect to r.f. voltage across the coil. If the stage is *grid* modulated or if the transmitter is keyed in a low level stage so that plate voltage appears on the tank at all times, then there is no point in connecting the rotor of the condenser to positive high voltage; simply by-passing it to ground with a high-voltage condenser will be sufficient.

If a tube has a plate-filament capacity of more than approximately 2 $\mu\mu$fd., it is desirable to connect across the section of the tuning condenser, to which the neutralizing condenser is connected, a capacitance exactly equal to the plate-filament capacity of the tube. The circuit will then be balanced regardless of the setting of the plate tank condenser, and neutralization will hold for all bands when once set for one of the higher frequency bands. In the 152-T amplifier, a small amount of capacity between the neutralizing end of the tank circuit and ground is obtained by mounting the tank condenser and coil vertically, the capacity between the bottom end of the tank circuit and the grounded chassis being approximately equal to the tube's plate-to-filament capacity. Besides providing this additional capacity, the vertically-mounted tank circuit also aids in keeping both the plate and neutralizing leads to a minimum length.

CHAPTER FOURTEEN

Speech and Modulation Equipment

THIS chapter deals with the design, construction, and operation of speech amplifiers and modulators, and with arrangements such as automatic modulation control circuits, which are normally a portion of the modulation equipment.

The audio equipment required in a 'phone transmitter will vary widely with different types of microphones, different modulation systems, and different amounts of power to be modulated. Since it would be virtually impossible to show designs that would be suited to any type of application, a number of good designs of conventional type will be shown to indicate the method of approach to the problem. These particular designs should more or less completely solve the speech amplifier problem in 75 per cent of the usual amateur transmitter installations. For those special cases where the designs shown are not completely suitable, small variations in the necessary respects will almost surely adapt the designs to individual needs.

The amplifiers and modulators shown have been thoroughly proven in actual use in amateur stations. Consequently, if these designs are followed exactly, no trouble will be experienced, either in getting them to work, or in their subsequent application to the job at hand. However, when making alterations in the designs to adapt the equipment to slightly different applications, due caution and forethought should be used in making the changes.

Hum Difficulties. It is more than likely that inductive hum pickup will be the problem most frequently encountered, both in making alterations in amplifier design, and in installing the speech equipment in the operating room or in the transmitter. The proximity of power supply equipment to the audio transformers or to the low-level grid leads should always be taken into consideration.

Any chokes or transformers in the low-level audio stages should be mounted as far as possible from power transformers and input filter chokes, which have relatively large surrounding a.c. fields. The audio transformers and coupling chokes can be properly oriented on the chassis before the holes are drilled for their mounting. A pair of headphones should be connected across the winding of each audio transformer or choke; 110-volts a.c. is then supplied to the primaries of all power transformers, and the audio transformer, or choke, is then rotated to determine the center of the hum "null." It should be bolted to the chassis in this position, even if it detracts from the neatness of the amplifier.

Some manufacturers offer special hum-bucking transformers for use in low-level audio stages; the transformers are so wound that they need not be specially oriented for minimum hum pickup.

Especial care need not be taken with high-level audio transformers, such as class B input and output transformers, if they are well-shielded and are not mounted too close to any power transformers.

The use of resistance coupling in the low-level audio stages of a speech amplifier makes it unnecessary to take precautions against inductive hum pickup. But grid and plate leads should be well isolated from power supply and high-level audio circuits, to prevent electrostatic pickup. However, it is usually much easier to shield the low-level section of a speech amplifier from electrostatic pickup than it is from inductive hum pickup such as can arise when transformer coupling is used.

A separate ground lead, from the speech amplifier to an external ground, is strongly advisable when the speech amplifier is not integral with the rest of the transmitter. With relay rack construction, in which the rack frame constitutes a common ground for both r.f. and audio units, a heavy copper bus run as direct as possible to a good external ground will suffice.

Amplifier Input Circuits. Various types of input circuits for speech amplifiers have been shown in the chapter *Radiotelephony Theory.* The majority of the speech amplifiers and modulators shown later on in this chapter have input circuits designed for the use of the dia-

Figure 1.

25-WATT AUDIO CHANNEL.

This speech amplifier or modulator unit will modulate any class C stage running from 35 to 50 watts input. Or it can be used to cathode modulate an amplifier stage running from 100 to 200 watts input. It comprises a high-gain speech amplifier working into a pair of pentode connected 42's or 6F6-G's in class AB with semi-fixed bias.

phragm-type crystal microphone. This type of input circuit has been shown, since that is the type of microphone most suited to amateur usage, and since the majority of amateurs now operating phone transmitters are using this type of microphone, or are contemplating the purchase of one. For those amateurs who have a particular preference for another type, such as the dynamic or the condenser, or for those who have another type microphone in good condition and who do not desire to purchase another, the special input circuits to the first speech stage shown in Chapter 8 can be adapted to the speech amplifiers to be described.

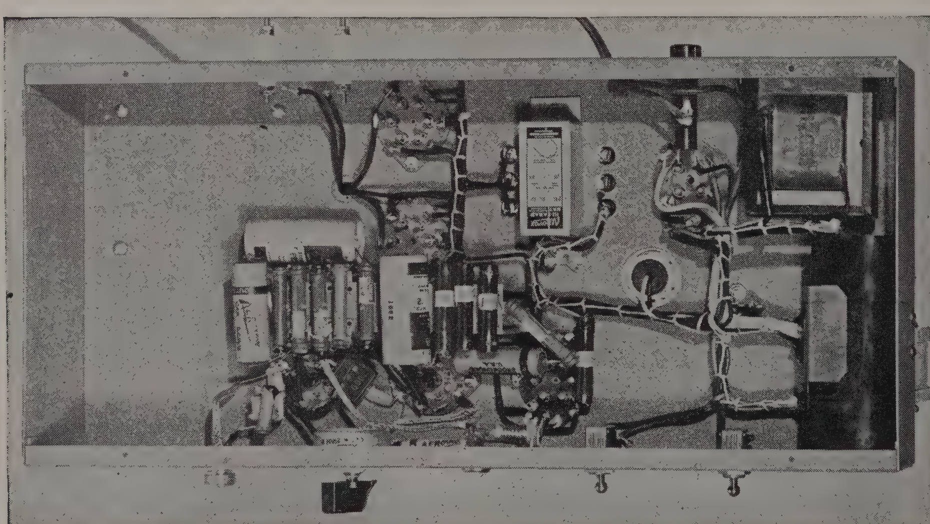
25-Watt General Purpose Modulator

For plate modulation, or combined plate and screen modulation of a low-powered transmitter, a modulator with an output in the vicinity of 25 watts is usually required. Such a unit is pictured in Figures 1 and 2, the schematic wiring diagram appearing in Figure

Figure 2.

UNDER-CHASSIS VIEW OF THE 25-WATT MODULATOR.

The filter choke and all resistors and condensers except C_{10} are mounted below the chassis. The use of a resistor strip adds to the appearance and facilitates checking and replacement of these units.



3. This modulator is simple and inexpensive to construct, and will plate modulate inputs of 40 to 60 watts on voice with excellent quality; the maximum output of the modulator, with voice frequency input and tolerable harmonic distortion, is about 25 watts.

While the unit can be used to drive a class B modulator, or to grid modulate a high powered grid modulated transmitter simply by tying a 15,000-ohm 10-watt resistor between the plates of the push-pull 42's, the unit is not recommended for such work, as it does not work as well into a variable load as do some of the other units described in this chapter. In other words, the unit works best when the output feeds into a constant load, such as when it is used to plate modulate a low-powered transmitter.

Tube Lineup. The first stage of the amplifier, a pentode-connected 6SJ7, is designed to operate from a crystal or other high impedance microphone. The input plug is of the shielded type, allowing a firm screw-on connection to the grounded side of the microphone cable.

A 6C5 in a conventional resistance coupled circuit amplifies the output of the 6SJ7 sufficiently to drive the triode-connected 42. The latter has more than sufficient output to swing the grids of the push-pull modulators with low distortion.

The values of the coupling condensers C_4 and C_6 were chosen with respect to R_5 and R_8 so that the gain will be attenuated in the extreme bass register (below 150 cycles). The advantages of bass suppression for voice transmission were covered in Chapter 8.

42's were chosen for use in the last two stages because they are inexpensive considering their power capabilities; also, they will give service under a moderate overload.

The output tubes are operated with semi-

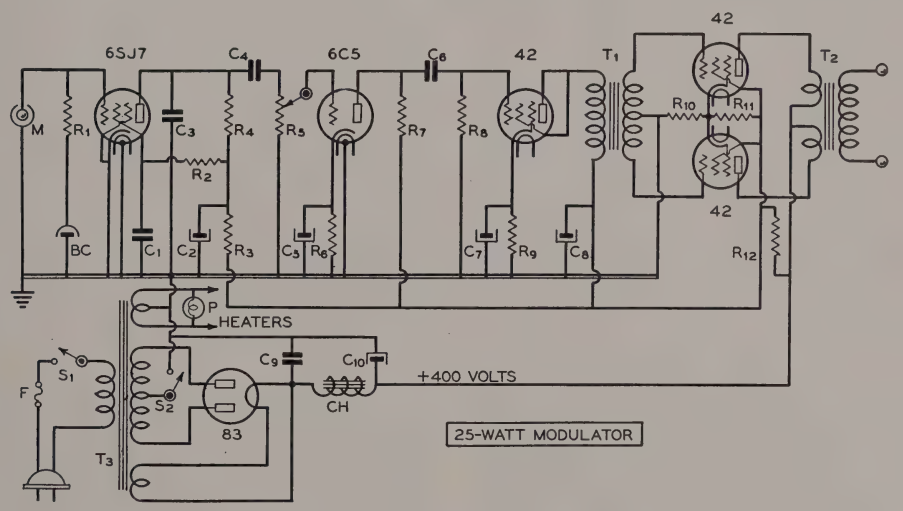


Figure 3.
WIRING DIAGRAM OF THE 25-WATT MODULATOR.

C_1—0.25-μfd. 400-volt tubular
C_2 — 12-μfd. 450-volt electrolytic
C_3—.0001-μfd. mica
C_4—.01-μfd. 400-volt tubular
C_5 — 10-μfd. 25-volt electrolytic
C_6—.01-μfd. 400-volt tubular
C_7 — 10-μfd. 50-volt electrolytic
C_8 — 12-μfd. 450-volt electrolytic
C_9 — 4-μfd. 600-volt paper
C_{10} — 8-μfd. 450-volt electrolytic
R_1—1 megohm, ½ watt
R_2—1 megohm, 1 watt
R_3—50,000 ohms, 1 watt
R_4—250,000 ohms, 1 watt
R_5—250,000-ohm potentiometer
R_6—2500 ohms, 1 watt
R_7 — 50,000 ohms, 1 watt
R_8 — 200,000 ohms, 1 watt
R_9 — 750 ohms, 10 watts
R_{10} — 250 ohms, 10 watts
R_{11}—10,000 ohms, 10 watts
R_{12} — 2000 ohms, 10 watts
T_1—3/1 pri.-to-½ sec. driver trans. (42 to 42's class A-B)
T_2—Variable ratio 30-watt output transformer
T_3—720 v., c.t., 125 ma.; 6.3 v., 4 a.; 5 v., 3 a.
CH—30-hy. 110-ma. choke (not over 200 ohms)
P—Pilot lamp
F—2-ampere fuse
BC—Bias cell
M — Shielded microphone connector

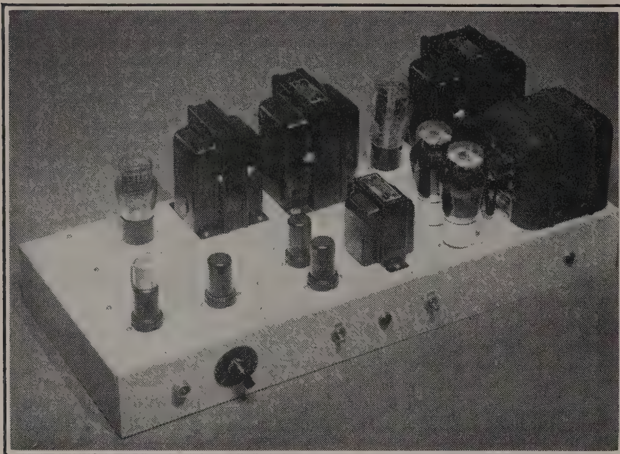


Figure 4.
TOP VIEW OF THE 60-WATT T-21 MODULATOR.
The power supply components are lined up along the rear half of the chassis, starting with the bias rectifier and ending with the oversize power transformer on the right rear. The audio frequency stages progress from left to right along the front of the chassis ending up with the multiple-match output transformer on the right end.

stabilized cathode bias. The resistor R_{11} stabilizes the screen voltage and the grid bias, and at the same time acts as a bleeder for the power supply.

Operation. The variable ratio output transformer makes the modulator adaptable to almost any transmitter. The taps should be connected so that a load of approximately 10,000 ohms, plate to plate, is placed on the 42's when the modulated stage is drawing normal plate current. The correct method of connecting the transformer taps for any particular installation can be determined quite easily by referring to the impedance ratio chart supplied with the particular make of transformer used.

As an example, if the modulated stage draws 100 ma. at 500 volts (such as a single 809), the load on the secondary of the modulation transformer will be 5000 ohms. Look up on the transformer chart the closest combination which reflects a 10,000-ohm plate-to-plate load on the primary when a 5000-ohm load is placed across the secondary of the transformer.

A 60-Watt T-21 Modulator Incorporating A.M.C.

The modulator illustrated in Figure 4 is designed primarily for use as a complete speech amplifier and modulator, to operate from a diaphragm-type crystal microphone, and to plate modulate about 150 watts input to a class C amplifier. It could, of course, also be used as a cathode modulator for 400 to 500 watts input to the stage; or, with about 20 db of feedback to the grids of the 6J5 drivers, it could be used as a high-level driver for a high-power class B stage.

A. M. C. Provision. Automatic modulation control has been incorporated into the design of the first stage of the amplifier. If it is desired to use the a.m.c. provision, and it is highly recommended that it be used, the a.m.c. rectifier may be coupled into the terminal marked "a.m.c. input." If it is not desired to use a.m.c., the terminal may be left open or grounded, as desired. Incidentally, there must be a biasing system incorporated into the a.m.c. rectifier, as shown in the one at the end of this chapter; some of the earlier a.m.c. systems had the biasing system incorporated into the speech amplifier, and, hence, did not need any bias on the a.m.c. rectifier. If an unbiased rectifier is used with this arrangement, the a.m.c. action will not come into effect until 100 per cent modulation is reached.

A New Phase-Inverter Circuit. The 6J5 phase inverter operates in a new-type circuit which is quite simple and yet which gives a reasonable amount of gain. On first glance it might appear that the 6J5 operates in the conventional "hot cathode" circuit, which has been used for some years with reasonable success. But, while the old circuit gave practically no gain in voltage in the phase inverter, by the changing of a few values and the addition of a resistor and a condenser, the voltage gain of the circuit has been increased to approximately 7 per side, or a total gain of about 14,—quite a worthwhile improvement for the addition of only one resistor and a condenser.

The operation of the circuit is simple enough, and should be apparent by inspection of the diagram. In the conventional arrangement, with C_7 and R_6 not in the circuit, when a voltage is impressed upon the grid of the 6J5, half the voltage output of the tube ap-

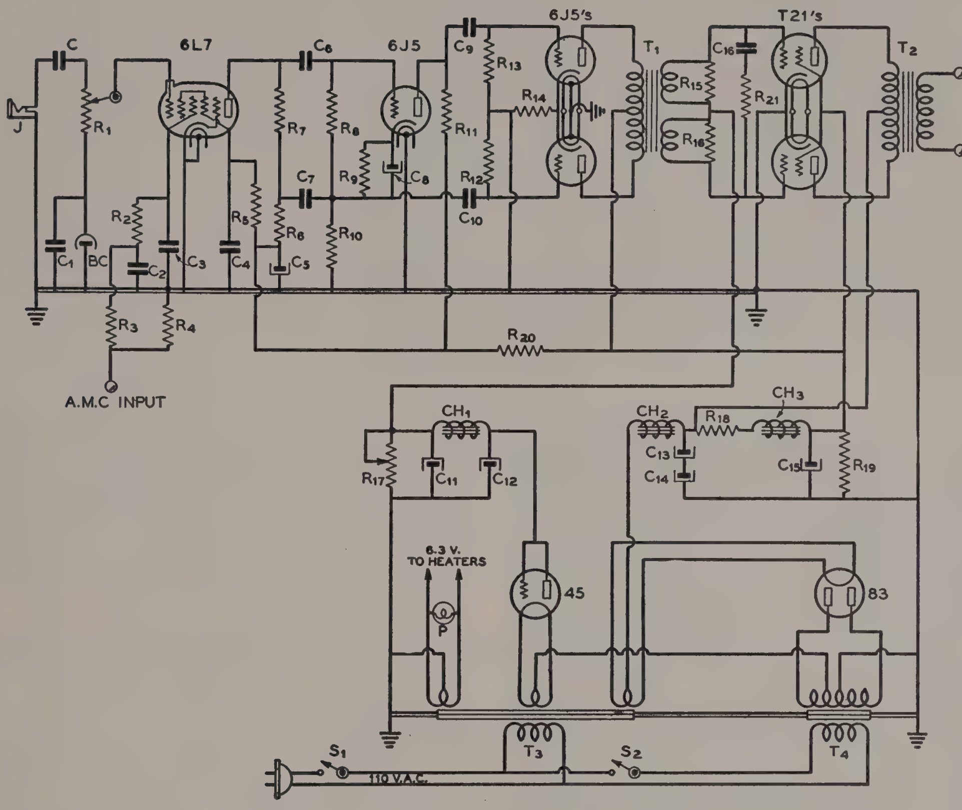


Figure 5.
WIRING DIAGRAM OF THE 60-WATT T-21 MODULATOR.

C, C_1—0.1-μfd. 400-volt tubular
C_2—.005-μfd. 400-volt tubular
C_3 — 0.5-μfd. 400-volt tubular
C_4 — 0.1-μfd. 400-volt tubular
C_5 — 8-μfd. 450-volt electrolytic
C_6—.01-μfd. 400-volt tubular
C_7 — 0.5-μfd. 400-volt tubular
C_8 — 10-μfd. 25-volt electrolytic
C_9, C_{10}—.01-μfd. 400-volt tubular
C_{11}, C_{12}—16-μfd. 100-volt electrolytic
C_{13}, C_{14}, C_{15} — 8-μfd. 450-volt electrolytic
C_{16}—.005-μfd. mica
R_1—1.0-megohm potentiometer
R_2—1.0 megohm, ½ watt
R_3—100,000 ohms, ½ watt
R_4—500,000 ohms, ½ watt
R_5 — 1.0 megohm, ½ watt
R_6—100,000 ohms, ½ watt
R_7, R_8—250,000 ohms, ½ watt
R_9 — 2500 ohms, ½ watt
R_{10}—100,000 ohms, ½ watt
R_{11}—50,000 ohms, ½ watt
R_{12}, R_{13}—250,000 ohms, ½ watt
R_{14}—600 ohms, 1 watt
R_{15}, R_{16}—10,000 ohms, 1½ watts
R_{17}—1500-ohm 10-watt slider or 1000-ohm 10-watt fixed
R_{18} — 2000 ohms, 10 watts
R_{19}—25,000 ohms, 20 watts
R_{20} — 5000 ohms, 10 watts
R_{21}—5000 ohms, 1½ watts
J—Microphone jack
BC—Bias cell
T_1—5:1 driver-to-6L6 trans.
T_2—Multi-match output trans.
T_3—2.5 v., 3.5 a.; 5 v., 3 a.; 6.3 v., 3 a. filament trans.
T_4—1030 v. c.t., 250 ma.; bias tap at 30 volts
CH_1—7.2-hy., 120-ma. choke
CH_2—13-hy., 250-ma. choke
CH_3—15-hy., 85 ma. choke
S_1—A.c. line switch
S_2—Plate on-off switch

pears across the cathode return resistor R_{10}. This voltage is fed back 180° out of phase with the incoming voltage, and in series with it. The resulting 50 per cent degenerative feedback reduces the gain of the stage to just more than 1. But, by isolating the cathode feedback voltage from the exciting voltage which appears across R_7, the plate circuit of the 6L7, the degenerative feedback is greatly reduced, and the stage attains almost normal gain,—in addition to its function as a phase inverter.

One consideration in the design is the shunt resistance of R_{10} and R_6, as compared to the resistance of R_{11}; (R_{10} and R_6 are effectively shunted as far as audio frequencies are concerned by the effects of condensers C_7 and C_5). The shunt effect of the first two should be equal to R_{11}. The most satisfactory way of obtaining this is to make R_{10} and R_6 each twice the value of R_{11}. This allows an equal audio voltage division between the plate and cathode circuits of the inverter tube. The plate impedance of the 6L7 is so high (approximately 0.8 megohm) as not to disturb the balance of the circuit materially.

P. P. 6J5 Drivers. The driver stage for the T-21's is perfectly conventional, and consists of a pair of 6J5's operating into a 5:1 driver-to-6L6 class AB_2 transformer. The swamping resistors, R_{15} and R_{16}, across the secondary of the driver transformer serve to improve the audio regulation. T-21's are used as the final modulator tubes, with 400 volts on their plates and operating with fixed bias.

Separate Bias Supply. The bias supply uses a 45 with the grid and plate strapped together, operating from the 30-volt tap on the power transformer. With the particular components that were used in the laboratory model of this modulator, when R_{17} was made a 1000-ohm 10-watt resistor, the bias voltage was the proper value on the T-21's. However, to allow for variations in tubes and equipment, it was felt best to specify a 1500-ohm adjustable resistor in this position. In any case, the shorting tap on the resistor will be very near to 1000 ohms.

The resistor-capacity network from grid to grid on the T-21's, C_{16} and R_{21}, was placed in the circuit to improve the waveform of the output at speech frequencies. Through the use of comparatively low value of coupling condensers from stage to stage within the amplifier, the frequency response of the amplifier drops quite sharply below 150 cycles. This reduces the difficulties resulting from hum pickup, and allows a higher relative modulation percentage to be obtained on the voice frequencies above 150 cycles, those that contribute most to the intelligibility.

The primary and secondary of the multimatch output transformer should be strapped

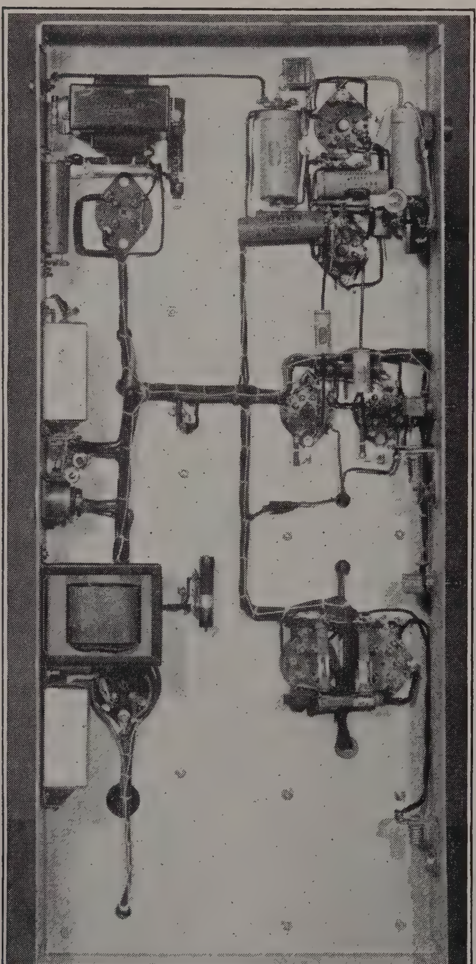


Figure 6.

UNDER-CHASSIS VIEW OF THE T-21 MODULATOR.

All power supply components are arranged along the rear side of the chassis and all audio components along the front. The single interconnecting cable between the two halves of the amplifier tends to minimize electrostatic coupling between them and hence to reduce hum pickup. However, to minimize electrostatic pickup from external sources it has been found desirable to place a metal cover on the bottom of the chassis. The chassis should be connected to external ground by an independent connection.

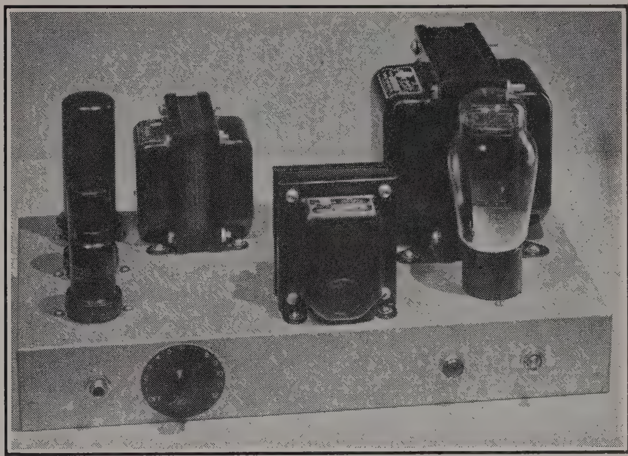


Figure 7.
FRONT VIEW OF THE 5-WATT AMPLIFIER.
The three audio tubes, the 6J5 first stage, the 6SJ7 second, and the 6L6 power amplifier are lined up along the left end of the chassis. The output transformer is alongside the 6L6; the other components are those associated with the power supply. The jack for the crystal microphone and the volume control are on the front drop of the chassis.

so as to present a plate-to-plate load of 4000 ohms to the T-21 tubes, with the value of secondary load into which the tubes are working. Maximum output and maximum modulating ability, with minimum harmonic distortion, will be obtained from the amplifier under these operating conditions.

5-Watt Speech Amplifier or Grid Modulator with Degenerative Feedback

Figure 7 illustrates a simple 5-watt amplifier specifically designed to operate from a crystal

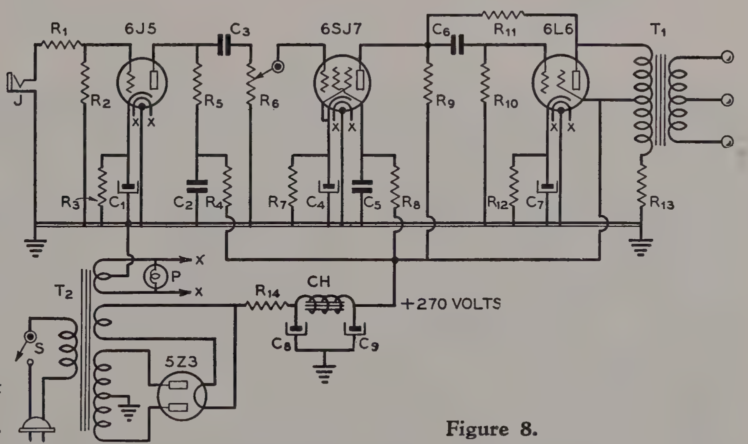


Figure 8.
WIRING DIAGRAM OF THE 6L6 GRID MODULATOR.

VALUES OF COMPONENTS

C_1—10-μfd. 25-volt electrolytic
C_2—8-μfd. 450-volt electrolytic
C_3 — .01-μfd. 400-volt tubular
C_4—10-μfd. 25-volt electrolytic
C_5 — 0.1-μfd. 400-volt tubular
C_6 — .05-μfd. 400-volt tubular
C_7—10-μfd. 25-volt electrolytic
C_8, C_9—8-μfd. 450-volt electrolytic
R_1 — 50,000 ohms, ½ watt
R_2 — 1.0 megohm, ½ watt
R_3—2000 ohms, ½ watt
R_4—25,000 ohms, 1 watt
R_5—250,000 ohms, ½ watt
R_6—500,000-ohm potentiometer
R_7—500 ohms, 1 watt
R_8—500,000 ohms, ½ watt
R_9—100,000 ohms, ½ watt
R_{10}—250,000 ohms, ½ watt
R_{11}—500,000 ohms, ½ watt
R_{12}—200 ohms, 10 watts
R_{13}—10,000 ohms, 10 watts
R_{14}—600 ohms, 10 watts
J—Microphone jack
P—6.3-volt pilot lamp
T_1 — Driver trans., p.p. 2A3's to 801's
T_2—720 v. c.t., 125 ma.; 5 v. 3 a.; 6.3 v. 4 a.
CH—25-hy. 140-ma. choke
S—A.c. line switch

microphone, and to be used as a grid modulator for a medium- to high-powered amplifier. A single-ended 6L6 is used as the output tube, with degenerative feedback from its plate back to its grid circuit. The use of degenerative feedback greatly lowers the plate impedance of the 6L6, and considerably reduces any harmonic distortion that might be introduced as a result of the operation of a single-ended beam tetrode stage. The reduction in the plate impedance of the 6L6 by feedback improves the regulation of the output voltage with respect to such changes in loading as are had when grid modulating an amplifier.

The Feedback Circuit. The addition of the single resistor R_{11} from the plate of the 6L6 back to the plate of the 6SJ7 amplifier stage, reduces the harmonic distortion, measured from the input of the 6SJ7 to the output of the amplifier, from approximately 11 per cent to less than 3 per cent at 5 watts output. The addition of the resistor for the shunt feedback circuit reduces the gain of the amplifier only a small percentage; there is ample gain to give full output when using a diaphragm-type crystal microphone on the input.

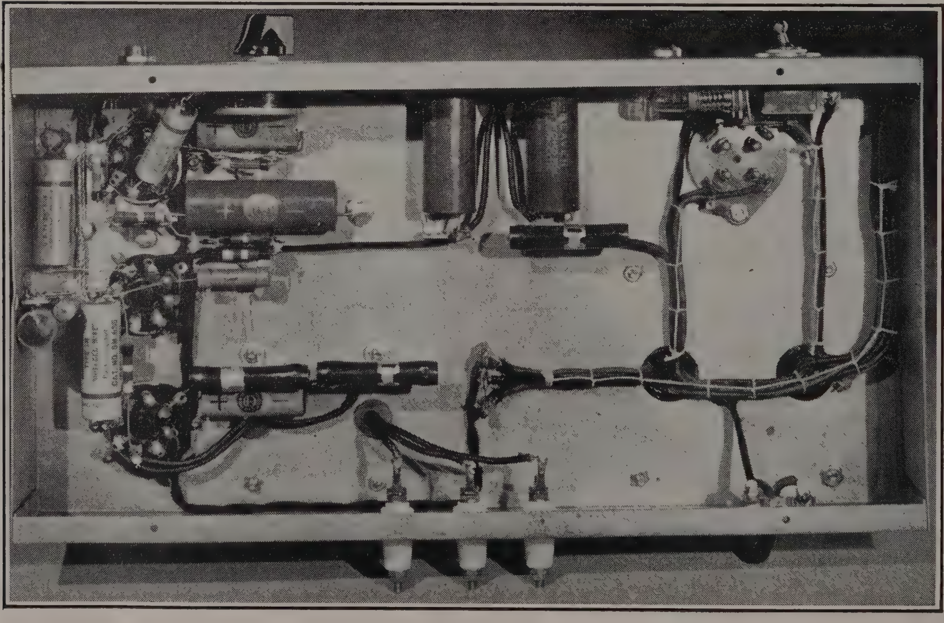
The power supply uses an input resistor instead of the more common input condenser or choke. The resistor serves to limit the voltage of the power supply to the proper value, both because of its action as a resistance, and because it acts as an input impedance ahead of the first condenser. It also contributes to the filtering action.

The Output Circuit. The shunt resistor, R_{13}, serves a triple purpose. In the first place, it acts as a load upon the output of the 6L6 to stabilize its output with respect to variations in load. Second, it acts as a bleeder upon the power supply to reduce the possibility of blowing the filter condensers in the interval between the heating up of the filament of the 5Z3 and the coming to operating temperature of the cathode of the 6L6. Third, its drain through the secondary of the output transformer opposes that of the 6L6, and tends to cancel the saturating action of the plate current of the 6L6 upon the core.

Figure 9.

UNDER-CHASSIS VIEW OF THE 6L6 AMPLIFIER OR GRID MODULATOR.

Under-chassis layout is comparatively simple and is made to present a neat appearance by cabling all the power supply leads. The three feedthrough insulators on the back-drop of the chassis are the three leads from the secondary of the modulation transformer; they can be used either to feed the grid return of the grid modulated stage, they may be used to plate modulate 10 to 15 watts input to a class C stage, or they may be fed to the grids of a medium power class B modulator.



The output transformer T_1 is a unit designed to be used as a driver transformer between push-pull 2A3's and the grids of a pair of 801's in class B. However, by using it in the amplifier as shown, it is possible to obtain a selection of four different impedance ratios from the plate of the 6L6 to the grid of the tube being modulated. If the 6L6 is fed into the side of the transformer originally meant for the 2A3's, the use of the full secondary will give a ratio of 2.35 to 1 step up; the use of half of the secondary will give about 1.2 to 1 step up. Then, if the 6L6 is operated into the side designed for the 801's, the use of the total secondary will give a ratio of 1.7 to 1 step up, and the use of half of the secondary will give a ratio of 1 to 0.85 step down. This latter ratio is the one most likely to be used when modulating medium-μ tubes at normal plate voltages. The step-up ratios would be used with medium-μ or low-μ tubes at comparatively high plate voltages and plate inputs up to 1 kilowatt.

If the output transformer is connected to the plate of the 6L6 in the manner for which it was designed (the 6L6 feeding into the 2A3 side), the secondary may be connected to the grids of a pair of medium-power class B tubes. Tests have shown that the amplifier thus connected has ample gain and power output to drive a pair of 809's, HY-25's, HY-40Z's, or a pair of 811's, HY-51Z's, TZ-40's at 1250 volts.

Push-Pull 2A3 Amplifier-Driver

A speech amplifier-driver for a medium-powered class B modulator is shown in Figures 10 and 11. The amplifier is designed to work out of a diaphragm-type crystal microphone, although any other type of input circuit could be used with equally good results. Alternative input circuits have been shown in Chapter 8.

The first stage utilizes one of the new single-ended metal pentodes: a 6SJ7. The gain control is between its plate circuit and the grid of the 6C5 second stage. The output

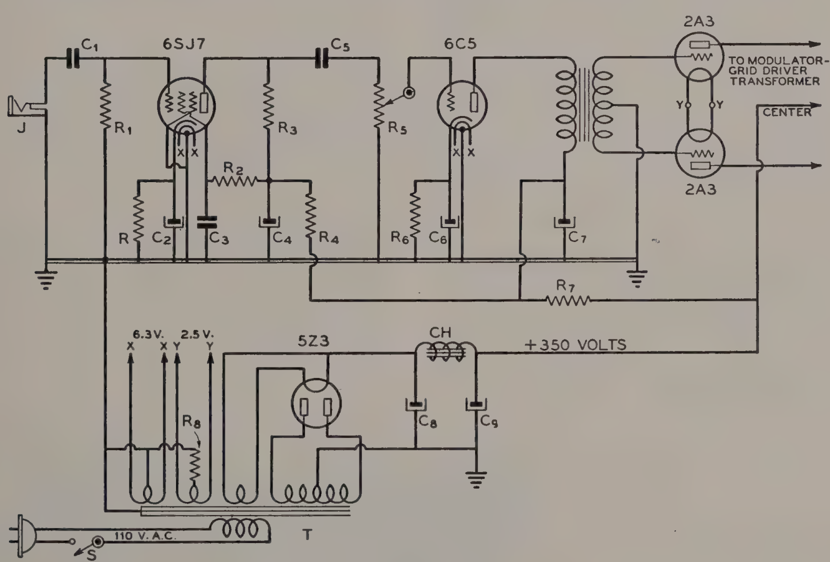


Figure 10.
WIRING DIAGRAM OF THE 2A3 AMPLIFIER-DRIVER.

- C_1 — .01-μfd. 400-volt tubular
- C_2 — 10-μfd. 25-volt electrolytic
- C_3 — 0.1-μfd. 400-volt tubular
- C_4 — 8-μfd. 450-volt electrolytic
- C_5 — .01-μfd. 400-volt tubular
- C_6—10-μfd. 25-volt electrolytic
- C_7, C_8, C_9—8-μfd. 450-volt electrolytic
- R—1000 ohms, ½ watt
- R_1 — 1.0 megohm, ½ watt
- R_2 — 1.0 megohm, ½ watt
- R_3—250,000 ohms, ½ watt
- R_4 — 10,000 ohms, ½ watt
- R_5 — 500,000-ohm potentiometer
- R_6 — 2500 ohms, ½ watt
- R_7 — 5000 ohms, 10 watts
- R_8—750 ohms, 10 watts
- P.p. input trans.—2:1 overall
- T — 700 v. c.t., 110 ma; 5 v., 3 a.; 2.5 v., 14 a.; 6.3 v., 4.5 a.
- CH — 25-hy. 140-ma. choke
- J—Microphone jack
- S—A.c. line switch

tubes are a pair of 2A3's, operating with a self-bias resistor in their common filament return. Operating in this manner, the 2A3's have an undistorted output of approximately 10 watts.

As a Driver. A pair of 2A3's operating in this manner will have ample output to drive almost any class B modulator whose output is 300 watts or less. The driver transformer for coupling the plates of the 2A3's to the grids of the class B stage is not shown, since it has been found best to have this transformer at the grids of the driven tubes, rather than at the plates of the drivers. The correct transformer step-down ratios for driving almost any class B tube have been set down in tabular form by the various transformer manufacturers. When the driver transformer is purchased, one should be obtained which has the proper ratio for the tubes to be used. Some manufacturers make multiple-ratio transformers which allow a proper match to be obtained for a large number of tubes.

A 3-wire shielded cable should be run from the output of the 2A3 tubes to the driver transformer at the grids of the class B tubes. This cable may be made any reasonable length up to 25 or 30 feet. Make sure that the insulation from the 3 wires to ground is ample to withstand about twice the d.c. voltage on the tubes.

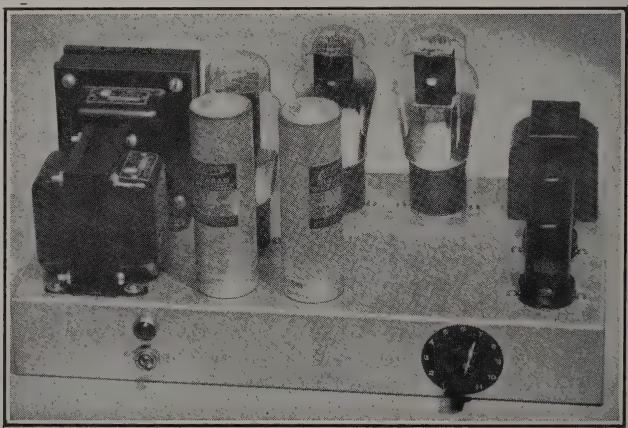
For driving a class B modulator of less than 75 watts output, type 45's may be substituted for the 2A3's with no changes in circuit constants. The 45's are less expensive.

10-Watt 6A3 Amplifier-Driver with Two Input Channels

An alternative speech or driver unit, which can serve additional duty in other jobs, is shown in Figures 12 and 13, and diagrammed in Figure 14. The unit features dual input channels, one high gain for a microphone, and one low gain for a phonograph pickup, with separate gain controls on each, main gain control which controls the gain of the mixed output of the input stages, and control of the output impedance between 8, 15, and 500 ohms by means of a selector switch. From the above information it is obvious that this amplifier is ideally suited for use as a public-address unit or as a recording amplifier, in addition to its suitability for use as a speech system. The amplifier may also be used as a driver for a class B modulator, provided the tubes driven require not more than 5 watts of audio. The reduction in the rated power output of the amplifier is due to the fact that an additional transformer must be used to couple the 500-ohm output of the amplifier to the grids of the class B tubes. The combined insertion losses of the transformer in the amplifier and the external 500-ohm-to-grid driver transformer will be in the vicinity of 3 db; hence, approximately half the output power of the 6A3 tubes will be lost.

The electrical design of the amplifier is more or less conventional in every respect. The two 6SJ7 input stages have common cathode, screen, and plate circuits. But the input of one tube is fed directly from a 1.0-megohm gain control for the microphone circuit (with a 50,000-ohm resistor in series with the grid of the tube, to reduce any possible tendency toward r.f. pickup), while the grid of the other is fed from a 20,000-ohm potentiometer in series with a 500,000-ohm resistor. The effect of this attenuator in the grid circuit of the second tube of the mixer is to reduce the gain of this stage, so that a conventional phonograph pickup will give substantially the same range of output as will a crystal microphone

Figure 11.
THE 2A3 SPEECH AMPLIFIER-DRIVER.
The jack for the crystal microphone is mounted on the right drop of the chassis directly alongside the grid lead of the 6SJ7 first speech stage. The low plate impedance of the 2A3's makes this amplifier an ideal driver for any medium power class B modulator.



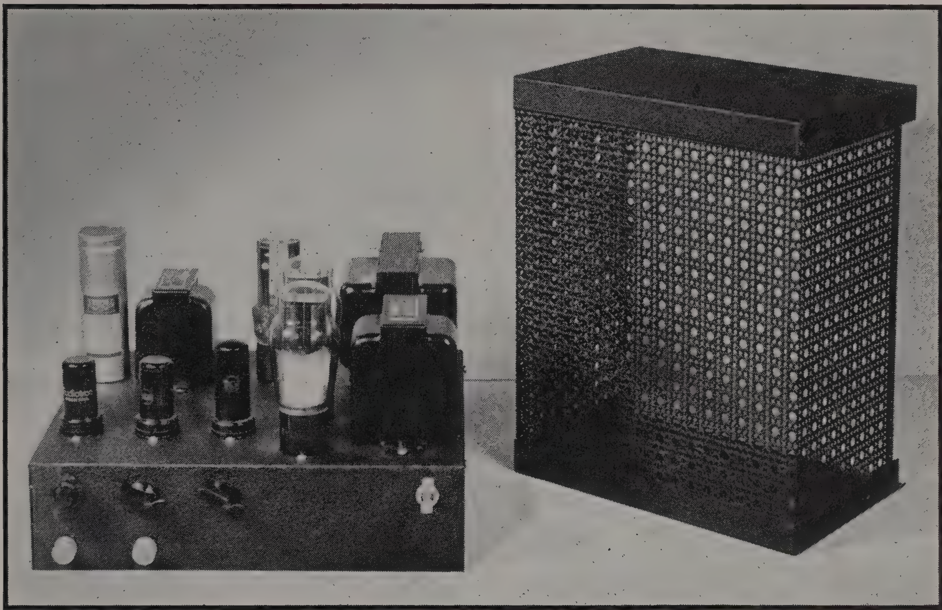


Figure 12.

FRONT VIEW OF THE DUAL-CHANNEL 6A3 AMPLIFIER.

Along the front drop of the chassis are the two input jacks with their individual gain controls directly above. The third knob is the main gain control for the amplifier. The output impedance selector switch is on the back drop of the chassis. The protective cover for the amplifier has been removed and is placed alongside.

on the other channel. The gain control in the grid circuit of the 6N7 controls the mixed output of the two 6SJ7 tubes.

The 6N7 phase-inverter circuit is the conventional "floating paraphase" circuit which has recently become so popular. Also conventional is the power supply and the 6A3 output stage. Since cathode bias is used on the 6A3's, their power output is limited to about 10 watts. The output transformer has taps for various values of load impedance; the desired impedance tap is selected by means of S_2.

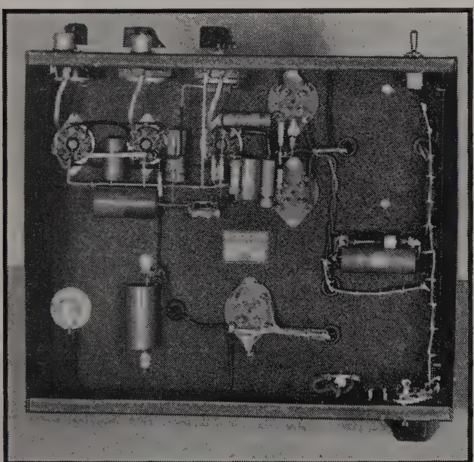


Figure 13.

UNDER-CHASSIS OF THE DUAL-INPUT 6A3 AMPLIFIER.

Note the simplicity of the design and the relatively small number of components required for this amplifier. The output impedance selector can be seen on the rear of the chassis.

Class B 809 Modulator

Figures 15 and 16 illustrate and show the schematic of a class B modulator using a pair of 809's. This modulator is designed to be driven by the push-pull 2A3 speech amplifier-driver shown above. A pair of 45's also could be used as drivers, but the 2A3's will have a reserve of driving power that will make for better quality from the modulator.

Voice Modulation Operation. With the 809's operating at 750 volts plate and 4.5

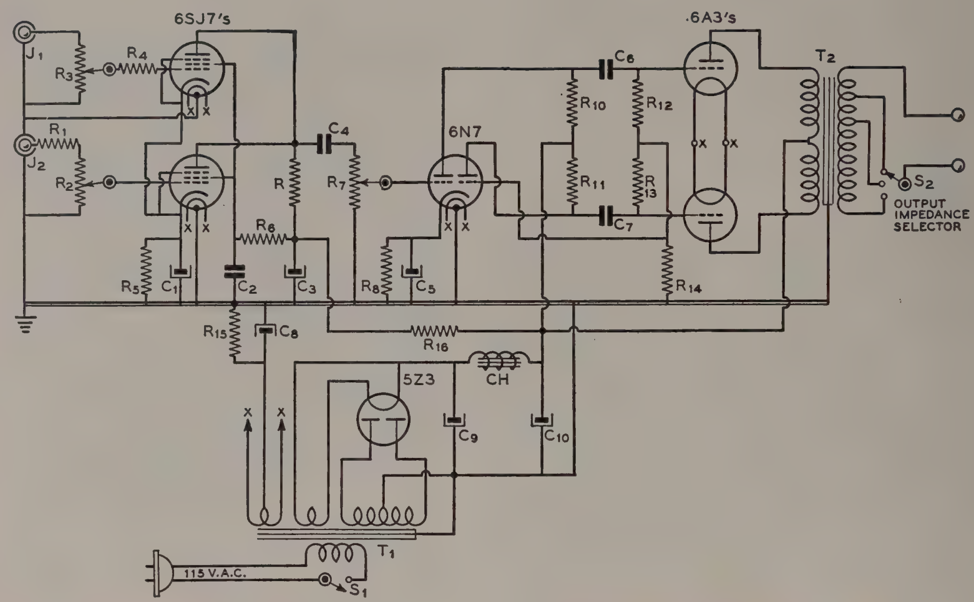


Figure 14.

WIRING DIAGRAM OF THE GENERAL PURPOSE 10-WATT 6A3 AMPLIFIER.

C_1 — 10-μfd. 25-volt electrolytic
C_2 — 0.1-μfd. 400-volt tubular
C_3 — 8-μfd. 450-volt electrolytic
C_4—.005-μfd. 400-volt tubular
C_5 — 25-μfd. 25-volt electrolytic
C_6, C_7—0.01-μfd. 400-volt tubular
C_8 — 50-μfd. 100-volt electrolytic
C_9 — 4-μfd. 600-volt oil-filled
C_{10} — 8-μfd. 450-volt electrolytic
R—250,000 ohms, ½ watt
R_1—500,000 ohms, ½ watt
R_2 — 20,000-ohm potentiometer
R_3 — 500,000-ohm potentiometer
R_4—50,000 ohms, ½ watt
R_5—800 ohms, ½ watt
R_6 — 1.0 megohm, ½ watt
R_7—500,000-ohm potentiometer
R_8 — 3000 ohms, ½ watt
R_{10}, R_{11}, R_{12}, R_{13} — 250,000 ohms, ½ watt
R_{14}—100,000 ohms, ½ watt
R_{15} — 800 ohms, 10 watts
R_{16}—50,000 ohms, ½ watt
T_1—700 ct. 120 ma., 5 v. 4 a., 6.3 v. 4.7 a.
T_2—5000 ohms pri. to 8, 15, 500 sec.
J_1 — Crystal microphone jack
J_2 — Phonograph input jack
S_1—A.c. on-off switch
S_2 — Output impedance selector

volts of bias, the plate-to-plate load should be 4800 ohms for maximum speech-waveform peak audio output. Under these conditions of operation, the instantaneous peak output from the tubes will be about 300 watts, which will allow the 809's to modulate an input of 300 watts to the class C amplifier. With 900 volts on the 809's, the proper plate-to-plate load resistance is 6200 ohms, and the peak output will be approximately 350 watts. If the plate voltage is raised to 1000 and the bias to 8 or 9 volts, the proper plate-to-plate load value is 7200 ohms, and the tubes will deliver a peak output of 400 watts, allowing them to voice modulate an input of 400 watts to the final stage.

Under all the above conditions of operation, full output from the 809's will be obtained when they are driven to an average plate current of approximately 160 ma. as indicated by the milliammeter M in the plate circuit. Testing of the modulator, with sine-wave audio as generated by an audio oscillator, is not to be recommended, except for a very short period of time, just long enough to make the measurement. If continuous modulation with a sine-wave tone is attempted, the maximum plate dissipation ratings of the 809's will be exceeded.

The input transformer ratio of total primary to half secondary should be approximately 4.5 to 1 for all conditions of operation.

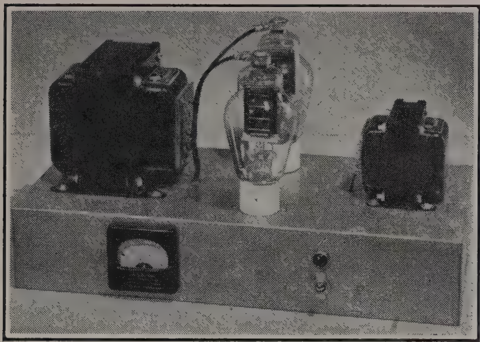


Figure 15.

CLASS B 809 MODULATOR.

Multiple ratio transformers have been used both in the grid and plate circuits of the modulator to increase its flexibility in matching various driver combinations and in coupling to various values of load impedance.

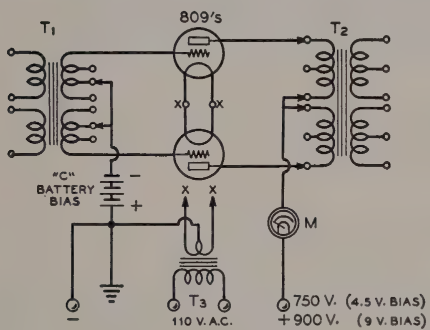


Figure 16.

SCHEMATIC DIAGRAM OF THE 809 MODULATOR.

T_1 — Multiple-ratio input transformers; 4.5:1 step-down ratio usually used
T_2 — Multiple-impedance output transformer
T_3 — 6.3-volt 5-ampere filament transformer
M—0-250 d.c. milliammeter

Sine-Wave Operation. If it is desired to operate the 809's under the sine-wave audio conditions for modulating a smaller input to the class C stage, the following conditions will apply: plate voltage, 500; grid bias, 0; plate-to-plate load impedance, 5200 ohms; power output, 60 watts (which will modulate an input of 120 watts to the class C stage); maximum signal plate current, 200 ma. Another set of conditions, recommended for somewhat greater power output with sine-wave audio, are: plate voltage, 750; grid bias, 4½; plate-to-plate load, 8400 ohms; power output, 100 watts (which will sine-wave modulate 200 watts input to the class C stage); maximum signal plate current, 200 ma. The correct driver transformer step-down ratio for these operating conditions is also 4.5: 1.

Alternative conditions for sine-wave operation of the 809 modulator are given under the ICAS ratings for the tube. With 1000 plate volts, 10 volts of grid bias, and a plate-to-plate load resistance of 11,600 ohms, the pair of 809's have a class B sine-wave rating of 145 watts. The tubes will, under these conditions, be able to plate modulate 290 watts into the class C amplifier, neglecting insertion losses in the modulation transformer.

150-Watt Class B 811 Speech System Incorporating Splatter Suppressor

Figure 17 shows a rear view of a complete speech amplifier and modulator, utilizing a pair of 811's in the class B output stage. The speech amplifier uses a 6SJ7, 6N7, and ends up in a pair of 6L6's, with degenerative feedback, as the driver for the class B tubes. Ample gain is afforded by the speech circuit for operation from a crystal microphone, or one of the new high-impedance dynamic types. The output circuit of the modulator incorporates a combination splatter suppressor and low-pass filter circuit. The rectifier tube, in series with the plate lead to the final amplifier, eliminates the negative plate current swings which can be caused by large negative modulation peaks, while the 3500-cycle cutoff low-pass filter prevents the transmission of high-frequency splatter components generated in the rectifier tube, and attenuates speech components or distortion falling above 3500 cycles.

Since this modulator is a portion of the 250-watt bandswitching 813 transmitter described in Chapter 16, the circuit diagram is not also shown here. The reader is referred to the chapter *Transmitter Construction* for the circuit diagram and further information on this complete speech and modulator system. The unit is capable of sine-wave modulating up to 300 watts input to the class C stage, with a 1250-volt plate supply for the 811's.

Speech-Modulator Unit with TZ-40's for 600 Watts Input

Illustrated in Figures 18 and 19, and diagrammed in Figure 20, is a complete speech channel capable of plate-modulating an in-

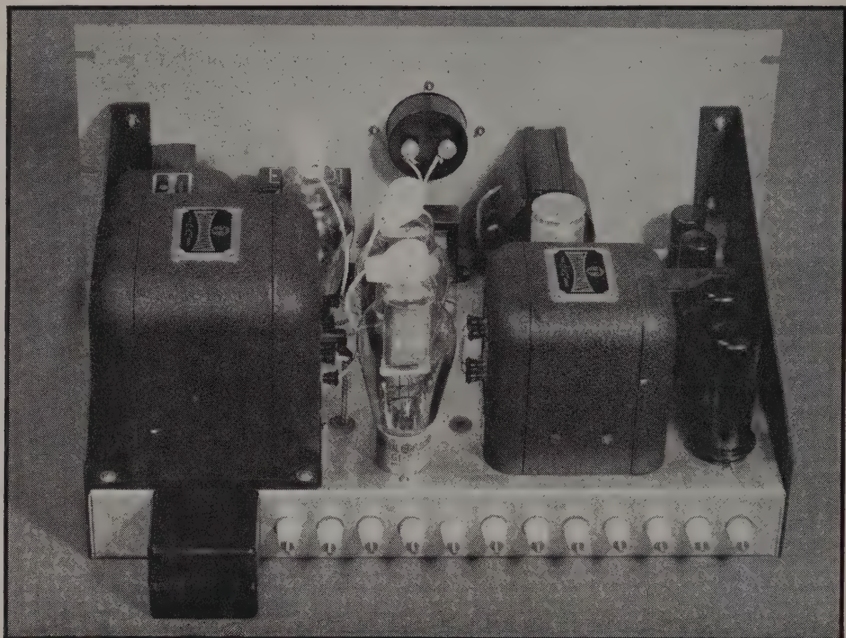


Figure 17.
REAR VIEW OF THE CLASS B 811 MODULATOR INCORPORATING SPLATTER SUPPRESSOR.

put of between 500 and 600 watts on voice. It incorporates a.m.c., inverse feedback, and other desirable modern features.

The combined speech amplifier and class B modulator, with the associated power supply for the speech amplifier, is built upon one 24 x 10 x 3 inch metal chassis. The underside of the chassis is not painted; the plated cadmium finish on this side facilitates the grounding of the various components.

The power supply for the speech stages is mounted along the left hand side of the chassis. Then there are mounted, in a row, the 6J7 first audio stage, the 6L7 a.m.c. amplifier, and the 6F6 last audio. Then, in the next row, in front, is the multitap driver transformer for the class B stage, then the two 6V6 drivers and, in back, the coupling transformer between the 6F6 and the two 6V6G's.

On the right hand end of the chassis are mounted the two TZ40 modulators and their associated class B output transformer.

Looking at the front of the chassis, at the extreme left, the on-off switch for all filaments and for the plate supply for the speech amplifier can be seen. The plate supply for the TZ40's is controlled at the transmitter proper. The next switch is the on-off switch for the a.m.c. circuit. Then comes the gain control, the microphone input jack, and the binding post for connection to the a.m.c. peak rectifier.

The under-chassis view is practically self-explanatory. At the extreme right end of the chassis is the 7.5-volt filament transformer for the TZ40's, and to the left of the center of the chassis are mounted the resistor plates. Only the upper one can be seen, as the two are mounted one above the other.

The speech amplifier uses a 6J7 metal tube connected as a high-gain pentode in the input. The circuit is conventional, and the tube is designed to operate from a diaphragm-type crystal microphone. The closed circuit jack on the input of the amplifier is shielded by a small metal can to eliminate any possibility of coupling between the output of the amplifier and the input circuit. Since the large metal spring of the jack is at grid potential, it is desirable to shield it from the output circuit of the 6V6G's, and from the a.m.c. lead which runs very close to the jack.

Automatic Modulation Control. The second stage of the amplifier—the a.m.c. stage—utilizes a 6L7 tube. The 500,000-ohm volume

control is placed between the plate circuit of the 6J7 and the control grid of the 6L7. It is important that this potentiometer be of the insulated-shaft type, since the entire 6L7 circuit operates considerably above ground potential.

The 879 reverse peak rectifier should be connected as follows: the plate of the tube should be connected directly to the a.m.c. binding post on the amplifier, and the filament of the tube should be connected to the lead that goes to the plates of the modulated class C amplifier. The filament should be lighted from a 2.5-volt filament transformer that is adequately insulated for twice the average plate voltage of the modulated amplifier, plus 1000 volts. Also, it is often a good idea to remove the negative peak rectifier as far as conveniently possible from both the speech amplifier and the class C final.

Since the injection grid of the 6L7 a.m.c. amplifier is 70 to 90 volts above ground potential (the whole a.m.c. stage is, as mentioned before, at this potential above ground), the 879 peak rectifier will begin to operate when the plate voltage on the class C amplifier becomes less than 70 or 90 volts, whatever the case may be. Then, as the modulator tends to drive the plate voltage lower than this, the gain on the speech amplifier will be reduced as the injector-grid bias on the 6L7 becomes negative. As this negative bias is increased, the signal output of the modulator is reduced. The final result: the output voltage of the modulator is reduced to an amount that will not cut the negative-peak plate voltage on the class C stage to zero; consequently, there is no overmodulation.

The gain on the speech amplifier may be run up to an amount which will permit a higher average voice level from the transmitter without any chance of overmodulation in any case. When the resulting signal is heard over the air, the transmitter seems to be modulated at a much higher percentage, although there is no tendency toward overmodulation splatter or hash.

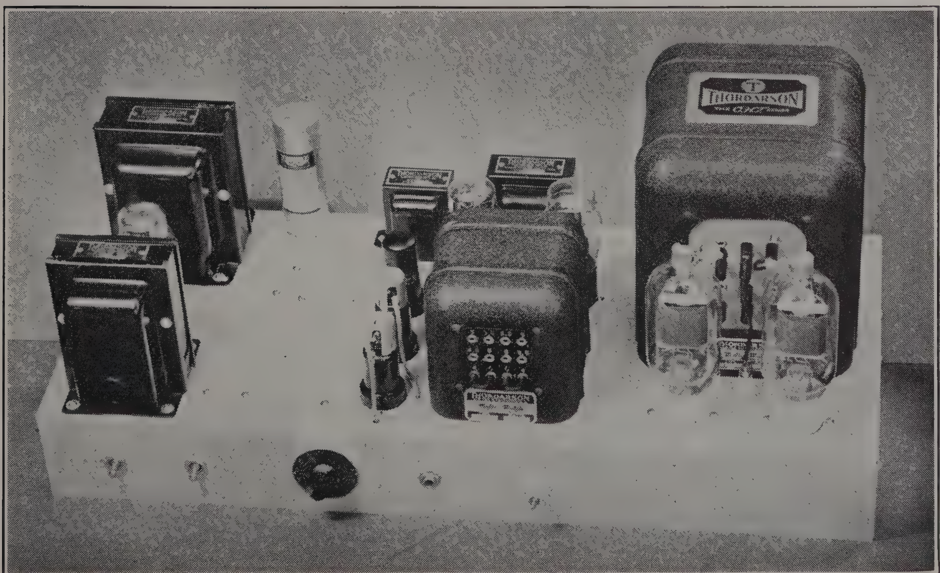
The 6V6G Drivers. A pair of 6V6's or 6V6G's are used as drivers for the TZ-40's. By using degenerative feedback from the secondary of the driver transformer to the screens of the 6V6's, the plate impedance of these tubes is lowered, thus making them well suited for use as drivers.

Beam tetrodes, when connected in the conventional manner, are not particularly well-suited as drivers for a class B stage unless a considerable amount of swamping is used. The high plate resistance of the tubes in the conventional method of connection causes a

Figure 18.

FRONT VIEW OF THE TZ-40 SPEECH AMPLIFIER-MODULATOR.

This combined speech amplifier and modulator will fully modulate up to 600 watts on voice. It incorporates inverse feedback, a.m.c., and other features.



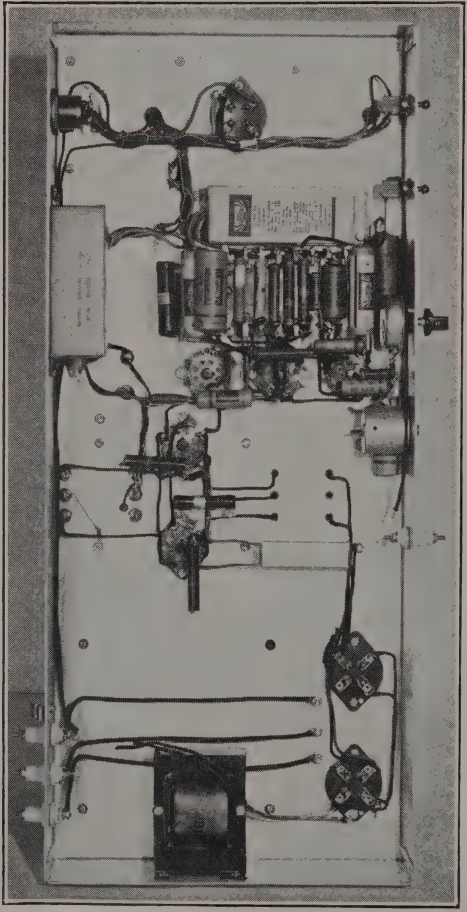
large drop in output voltage when any increase in load is placed upon them.

When first placing the amplifier in operation, it is very important that the screens be connected to the proper side of the class B modulation transformer secondary. The only way of finding out which side is the proper one is to connect up the amplifier and try it out. It is best not to have the plate voltage on the TZ40's when this test is made; something may flash over. Should the 6V6G's oscillate, reverse the connections between the screen grid coupling condensers and the class B grids, and the correct phase relation between the screen and plate voltages will be obtained.

Figure 19.

UNDER-CHASSIS VIEW OF THE TZ-40 MODULATOR.

The use of a resistor terminal plate, under-chassis wiring, and placement of components can be seen.



TZ-40 Operating Conditions. The TZ40's operate with zero bias under the conditions recommended by the manufacturers. The standing plate current on the two tubes is approximately 45 ma. with an applied plate voltage of 1000 volts. It will be somewhat higher, in the vicinity of 60 ma., if the full rated plate voltage of 1250 volts is used. Since this value of standing plate current results in an appreciable amount of plate dissipation, a small amount of grid bias is desirable, in order to lower the plate current under no-signal conditions. A pair of $4\frac{1}{2}$-volt batteries in series to give 9 volts is suitable as bias for 1250-volt operation.

For maximum peak power output from the TZ40's (for the adjustment which will modulate the greatest class C input with voice), the plate-to-plate load impedance for the 1000-volt conditions would be 5100 ohms. Under these conditions of operation, the modulator would be capable of 100 per cent voice-modulating at input of 500 watts to the class C stage; the plate current on the TZ40's should kick up to 200 to 250 ma. under normal modulation.

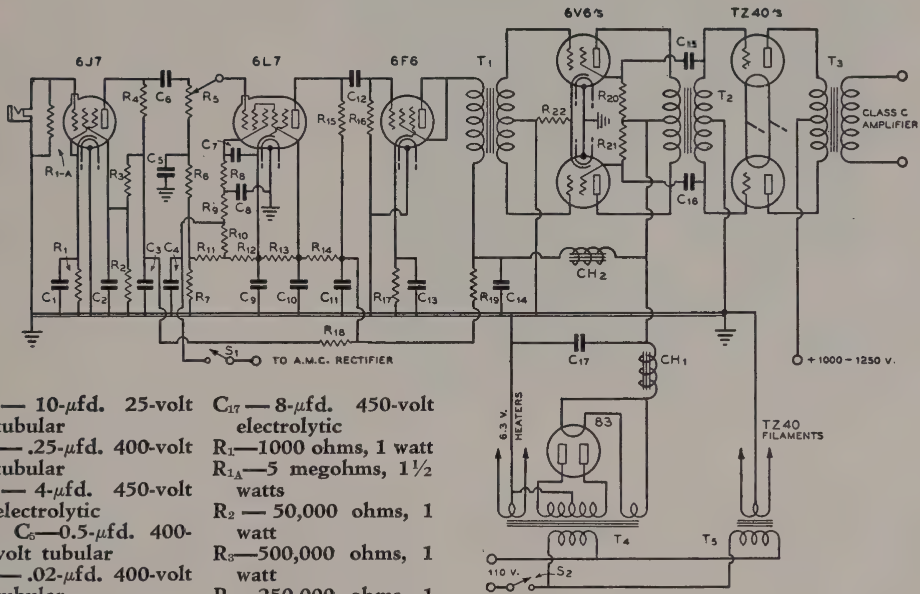
For maximum peak modulating capabilities at 1250 volts, the plate-to-plate load value should be 7400 ohms; the unit would be capable of fully modulating 600-watts input, and the plate current would kick up to 175 to 225 ma. under full modulation.

If it is desired to operate the class B stage under the conventional conditions for maximum *sine-wave* audio ouput, the plate-to-plate load resistance would be 6800 ohms under the 1000-volt conditions; the power output would be 175 rated watts, and the plate current would kick up to 250 to 275 ma. on peaks.

Complete 203Z Modulator and Speech Amplifier for Inputs Up to 800 Watts

Figures 21 and 22 show a speech amplifier and class B modulator suitable for modulating inputs from 400 to 800 watts input to the class C final stage. The speech amplifier portion of the modulator is more or less conventional, except for the inclusion of automatic peak compression to allow a higher average percentage of modulation without the danger of overmodulation on occasional loud voice peaks. The delay action in the compressor (the percentage of modulation at which compression starts) can be controlled by means of the potentiometer R_{14}. All components in the 6J7 first speech stage should be thor-

Figure 20.
SCHEMATIC DIAGRAM OF THE TZ-40 MODULATOR AND ASSOCIATED A.M.C. SPEECH AMPLIFIER.



C_1 — 10-μfd. 25-volt tubular
C_2 — .25-μfd. 400-volt tubular
C_3 — 4-μfd. 450-volt electrolytic
C_4, C_5—0.5-μfd. 400-volt tubular
C_6 — .02-μfd. 400-volt tubular
C_7 — 0.1-μfd. 400-volt tubular
C_8—.002-μfd. 400-volt tubular
C_9 — 8-μfd. 450-volt electrolytic
C_{10}—0.5-μfd. 400-volt tubular
C_{11} — 8-μfd. 450-volt electrolytic
C_{12}—.05-μfd. 400-volt tubular
C_{13} — 10-μfd. 25-volt tubular
C_{14} — 8-μfd. 450-volt electrolytic
C_{15}, C_{16} — 8-μfd. 450-volt electrolytic
C_{17} — 8-μfd. 450-volt electrolytic
R_1—1000 ohms, 1 watt
R_{1A}—5 megohms, 1½ watts
R_2 — 50,000 ohms, 1 watt
R_3—500,000 ohms, 1 watt
R_4—250,000 ohms, 1 watt
R_5—500,000-ohm potentiometer
R_6—500,000 ohms, 1 watt
R_7 — 4500 ohms, 5 watts
R_8—1 megohm, 1 watt
R_9—100,000 ohms, 1 watt
R_{10}—500,000 ohms, 1 watt
R_{11}—350 ohms, 1 watt
R_{12}—150 ohms, 1 watt
R_{13} — 5000 ohms, 5 watts
R_{14} — 7500 ohms, 5 watts
R_{15}—100,000 ohms, 1 watt
R_{16}—100,000 ohms, 1 watt
R_{17} — 750 ohms, 10 watts
R_{18} — 10,000 ohms, 5 watts
R_{19} — 2000 ohms, 5 watts
R_{20}, R_{21}—5000 ohms, 3 watts
R_{22} — 300 ohms, 10 watts
T_1 — Triode power tube to p.p. power tube driver transformer
T_2—Multi-match class B input transformer
T_3—Multi-match class B output (300 watt)
T_4—745 c.t., 145 ma.; 5 v. 3 a.; 6.3 v., 4.5 a.
T_5 — 7.5 volts, 4 amperes
CH_1—10-hy., 150-ma. filter choke
CH_2 — 10-hy., 65-ma. filter choke
S_1 — A.m.c. on-off switch
S_2—110-v. a.c. switch

oughly shielded to prevent grid hum, and to reduce the possibility of either r.f. or audio feedback.

Operation of the Class B 203Z's. The class B operating conditions recommended by the manufacturer for sine-wave audio output are 7900 ohms plate to plate at 1250 volts on the plate and 4½ volts of grid bias. Under these conditions the tubes will deliver 300 watts of sine-wave audio. For maximum speech audio output the plate-to-plate load resistance should be reduced to 5500 ohms. Under these conditions, the tubes will modulate an input of 800 watts, as compared to the 600 watts they will modulate under sine-wave audio operating conditions.

Power supplies, both for the speech amplifier portion and for the class B stage, are external. 1250 volts will be required for the 203Z's, and about 350 volts for the speech

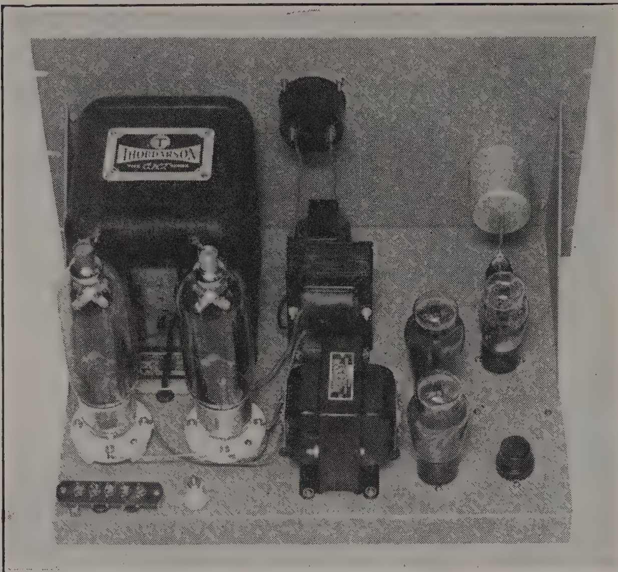


Figure 21.
203-Z SPEECH-MODULATOR FOR INPUTS TO 800 WATTS.
The complete speech amplifier and class B output stage are built upon a single relay-rack panel and its associated chassis. The speech amplifier incorporates automatic peak limiting and uses a pair of 2A3's as drivers for the 203-Z's.

amplifier portion. The 1250-volt supply should have good regulation up to a maximum drain of 350 ma., and the 350-volt supply should be capable of handling 125 ma. continuously.

Simplified Automatic Modulation Control

Figure 23 shows the circuit of a simplified method of obtaining the necessary bias required for an automatic-modulation-control system. This rectifier circuit must be used with the 60-watt T-21 modulator shown earlier in this chapter, if satisfactory a.m.c. action is desired. Through the use of the circuit illustrated, the bias required for all a.m.c. systems is placed on the rectifier tube itself, instead of being placed on the cathode of the a.m.c. tube in the speech amplifier. This greatly simplifies the design of the a.m.c. stage in the speech system.

"Advance" Bias System. In the circuit diagram, this "advance" bias is obtained by means of a voltage divider, consisting of a 50,000-ohm and a 500,000-ohm resistor, which reduces the d.c. plate voltage applied to the diode cathode about 9 per cent. This acts as the "advance" bias. The resistor R_1 can be of the 1-watt size for plate supplies up to 1000 volts, and a 2-watt for up to 2000 volts. The 500,000-ohm resistor can be made of ten similar carbon resistors, wired in series and well insulated from the chassis. C_1, R_1, and R_2 can be mounted on bakelite resistor mounting strips or panels about 1 inch away from the chassis, with the strip mounted on stand off insulators. The diode filament transformer must also be well insulated between windings in order to withstand the peaks in the positive direction.

The Rectifier Diode. The diode itself must have sufficient inverse peak rating, which means that an 866 Jr. is suitable for use in sets with plate supplies up to 1000 volts, an 866 up to 2500 volts, and an 879 for higher plate supplies. Mercury vapor in the rectifiers seems to make no difference in operation at the low currents used in a.m.c. circuits.

The purpose of C_1 in the circuit diagram is to by-pass the audio peak overload voltage into the diode cathode. The diode then has the full amount of a.c. peak across it, and a little over 90 per cent of the d.c. plate voltage. C_1 can be a 0.5- or 1-μfd. 400- or 600-volt paper condenser, as long as it is mounted well in the clear of nearby grounds.

The control bias is developed across R_3, which can be of any value between 100,000 and 250,000 ohms. No condenser should be connected across this resistor unless there is some stray r.f. present. If there should be any, it must be by-passed with a small .002-μfd. condenser. The time delay circuit should be confined mainly to C_3 and R_5, which can have values of 0.5 μfd. and 1 megohm in most speech transmitters. Additional audio filter in the form of C_2, 0.1 μfd., and R_4, 0.5 megohm, is generally necessary to prevent audio feedback and a "blurbing" effect on high levels

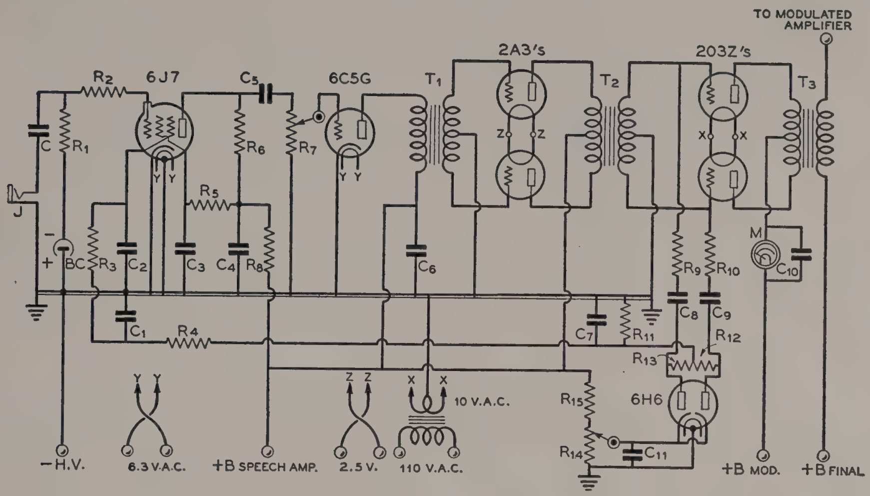


Figure 22.

WIRING DIAGRAM OF THE CLASS B 203-Z MODULATOR.

C — .01-μfd. 400-volt tubular
C_1 — 0.1-μfd. 400-volt tubular
C_2 — 0.1-μfd. 400-volt tubular
C_3 — .25-μfd. 400-volt tubular
C_4 — 0.5-μfd. 400-volt tubular
C_5 — .01-μfd. 400-volt tubular
C_6 — 0.5-μfd. 400-volt tubular
C_7—0.25-μfd. 400-volt tubular
C_8, C_9—0.1-μfd. 400-volt tubular
C_{10}—.002-μfd. mica
C_{11} — 1.0-μfd. paper, 400 volts
R_1—1.0 megohm, ½ watt
R_2—50,000 ohms, ½ watt
R_3—250,000 ohms, ½ watt
R_4—300,000 ohms, ½ watt
R_5 — 1.0 megohm, ½ watt
R_6—250,000 ohms, ½ watt
R_7 — 1.0 megohm potentiometer
R_8—50,000 ohms, ½ watt
R_9, R_{10}—2.0 megohms, ½ watt
R_{11}, R_{12}, R_{13}—100,000 ohms, 1 watt
R_{14} — 50,000-ohm potentiometer
R_{15}—100,000 ohms, 1 watt
J—Microphone jack
BC—Bias cell
T_1 — Push-pull input trans.
T_2—Class B input for 203Z's
T_3—300-watt variable-ratio modulation trans.
M—0-500 d.c. milliammeter

of speech input. These resistors can be of 0.5- or 1-watt size.

A.M.C. Tubes. It is possible to supply a.m.c. voltage to the control grid of an amplifier, such as to a 6K7 or even a 6N7. The suppressor grid of a 6C6, 6J7, or 6K7 requires about twice as much negative bias for the same reduction in gain as does the injector grid of a 6L7. It is advisable to use a 6L7 whenever possible. However, this a.m.c. circuit can be applied to nearly any existing phone transmitter, with hardly any changes in the speech amplifier.

A.M.C. Advantages. A.m.c. practically eliminates sideband splatter in all cases, and prevents modulation in excess of 100 per cent. In addition, it allows an average higher level of modulation, which results in better signal at the receiver. The two phone transmitters of the same carrier output, one with a.m.c. and one without, both not overmodulated, will have about 2 to 3 db difference in level. The 3 db increase available from the use of a.m.c. is equivalent to doubling the carrier signal in effect.

One other point should be mentioned; a.m.c. will handle only from 15 to 20 db excessive level peaks without considerable audio distortion. So don't try to push the average modulation level up to 99 per cent at all times. Use the manual gain control, too, and keep the level of modulation down to a point where it sounds right in a monitor. An oscilloscope will usually indicate 100 per cent modulation many times a minute on an average speech, when the gain adjustment is correct for good monitor quality.

The a.m.c. circuit shown in Figure 23, however, is *not* suitable for use with the TZ40 speech amplifier-modulator shown in Figure

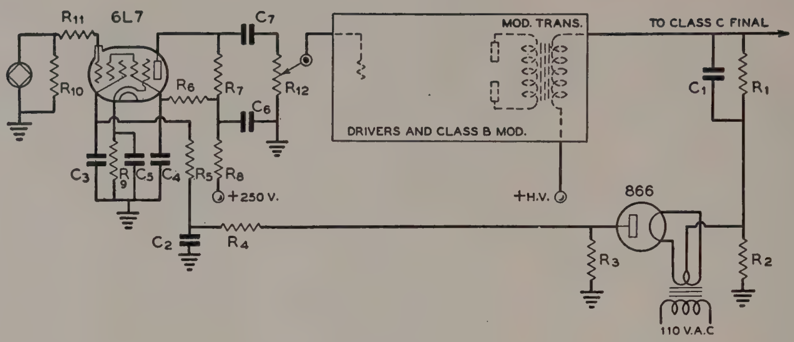


Figure 23.
A.M.C. ARRANGEMENT WITH SIMPLIFIED BIAS SYSTEM.

C_1 — 0.5-μfd. 600-volt tubular
C_2 — 0.1-μfd. 400-volt tubular
C_3, C_4—0.5-μfd. 400-volt tubular
C_5 — 10-μfd. 25-volt electrolytic shunted by .01-μfd. 400-volt tubular
C_6 — 0.5-μfd. 400-volt tubular
C_7 — .02-μfd. 400-volt tubular
R_1—50,000 ohms, 2 to 20 watts (see text)
R_2—500,000 ohms, 10 watts
R_3—100,000 ohms, 1 watt
R_4—500,000 ohms, ½ watt
R_5 — 1.0 megohm, ½ watt
R_6—250,000 ohms, 1 watt
R_7 — 200,000 ohms, 1 watt
R_8 — 30,000 ohms, 1 watt
R_9 — 1000 ohms, ½ watt
R_{10}—1.0 megohm, ½ watt
R_{11}—25,000 ohms, ½ watt
R_{12}—500,000-ohm potentiometer

18. All this speech amplifier requires is a half-wave rectifier. The same voltage ratings apply for this rectifier as for the one just described, with the c.t. of its filament connected directly to the plate voltage lead to the plate modulated stage, and the plate connected to the input terminal on the amplifier.

Efficient Splatter Suppressor

Phone splatter (adjacent channel interference) can be greatly reduced in a *plate modulated* transmitter by the incorporation of the circuit shown in Figure 24.

Simply inserting a low pass filter of suitable cut off frequency between the modulator and class C modulated amplifier will minimize splatter from high order harmonics generated in the modulator itself, but will do nothing towards eliminating splatter generated in the class C stage as a result of the negative peak clipping which occurs each time the plate voltage swings below zero.

By inserting a high vacuum rectifier between the modulator and low pass filter, negative peak clipping is virtually eliminated. A 5Z3 will be suitable for a d.c. plate voltage up to 2000 volts and d.c. plate current up to 300 ma. For greater plate current, two 5Z3's may be paralleled.

The 5Z3 filament transformer secondary must be insulated for at least twice the plate voltage, and should have reasonably low capacity to the primary and core.

For voice work, the cut off frequency of the low pass filter should be made about 3000 cycles.

Trouble Shooting in the Speech Amplifier

Great care is necessary in the design of speech amplifiers in order to prevent hum, distortion, and feedback at radio- or audio-frequencies. Certain precautions can be taken in building the speech amplifier, as related here: (1) Shield all low-level grid and plate leads. (2) Avoid overheating the shielded wires (rubber insulation) when soldering ground connections to the shield. (3) Shield all input and microphone connections. (4) Wire the filaments with twisted conductors. (5) Mount resistors and condensers as near as possible to socket terminals. (6) Orient the input and low-level audio transformers in a position of minimum hum when a.c. power is applied to the primaries of the power supply transformers. (7) Shield the input and low-level stage tubes. (8) Use a good ground connection to the metal

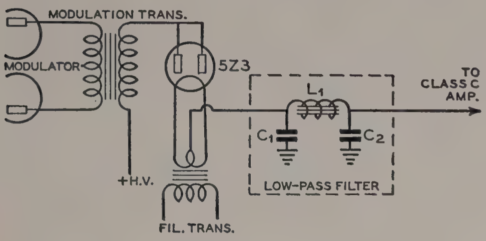


Figure 24.

SPLATTER SUPPRESSOR CIRCUIT.

C_1, C_2, L_1—Components of low-pass filter: characteristic impedance is the same as the class C load impedance; cutoff frequency is 3000 cycles.

chassis (waterpipe or ground rod connection). (9) Ground all transformer and choke coil cores. (10) Use metal cabinets and chassis, rather than breadboard construction. (11) By-pass low-level audio stage cathode by-pass electrolytic condensers with a .002-μfd. mica condenser, for the purpose of preventing rectification of stray r.f. energy which will sometimes produce hum.

The power supply for a speech amplifier should be exceptionally well filtered. This may require 3 sections of filter, consisting of 3 high-capacity condensers and 2 or 3 filter chokes. When space permits, the power supply should be placed several feet from the speech amplifier.

Shielding. The speech amplifier and microphone leads should be completely shielded for the elimination of r.f. feedback. A concentric or a *balanced* 2-wire r.f. transmission line to a remotely located antenna is the most effective method of preventing r.f. feedback into the microphone or speech amplifier circuits in the range of from 5 to 20 meters.

The impedance of ground leads at such short wavelengths makes it impossible completely to eliminate stray r.f. currents. End-fed antennas and single-wire fed systems are particularly troublesome with respect to r.f. feedback.

Audio feedback may cause motor-boating, whistling, or howling noises in the audio amplifiers. Insufficient by-pass capacity across the plate supply of a multistage speech amplifier is an additional cause of motor-boating. The first stage of a speech amplifier should have a resistance filter in its plate supply lead, which may consist of a 10,000- to 50,000-ohm 1-watt resistor in series with the positive B lead, with a 0.5-μfd. condenser connected to ground from the amplifier side of the series resistor. (See Figure 25.)

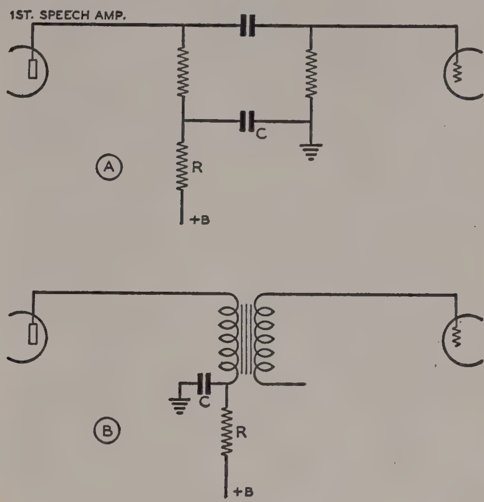


Figure 25.

RC FILTER CIRCUITS FOR USE IN DECOUPLING AUDIO STAGES.

The value of resistor R can be from 2000 to 50,000 ohms; C can be from 1 to 8 microfarads.

A defective tube will introduce hum or distortion, as well as affect the overall gain or power output of an audio amplifier. Incorrect bias on any amplifier stage will produce harmonic distortion, which changes the quality of speech. This bias voltage should be of the correct value for the actual plate-to-cathode voltage, rather than the plate supply output voltage; (these may be widely different in a resistance-coupled stage). Excessive audio input to any amplifier stage will produce amplitude distortion. Incorrect plate coupling impedances or resistances will cause distortion. A damaged or inferior microphone is another source of distortion. Cathode resistors should be by-passed with ample capacity to provide a low impedance path for the lowest frequencies. Push-pull, and especially class B amplifiers, require balanced tubes.

Power Supplies for Radiotelephony

A power supply for a radiotelephone transmitter should furnish nonpulsating d.c. voltage to the crystal oscillator or other source of frequency control. The amount of pulsation, or ripple voltage, should be less than 1 per cent of the d.c. voltage, especially for radio transmitters operating on very high frequencies. Hum or ripple voltage in the plate supply to the oscillator will frequency-modu-

late the r.f. output slightly. Each frequency multiplier stage increases the frequency modulation, until the carrier hum becomes objectionable in high-frequency transmitters. Many amateur 10-meter phones suffer from this difficulty, noticed especially with selective receivers.

The power supply for the front end of the speech channel must be thoroughly filtered, in order to avoid amplification of the ripple in the succeeding audio or speech amplifier stages. The plate supply for the final audio amplifier stage does not require as much filter as the preceding stages, and, in the case of a push-pull audio or driver stage, a single-section filter will suffice.

Buffer stages of a control-grid modulated transmitter must have very well-filtered plate supplies (more than the buffers in a plate-modulated transmitter), in order to prevent hum modulation in the grid circuit on which the speech audio frequencies are impressed. On the other hand, the plate supply for the grid-modulated stage itself does not require quite as much filter as does a comparable plate-modulated stage. This indicates that a single-section filter will suffice for a grid-modulated stage, whereas a 2-section filter is desirable for plate modulation. In the event that only a single-section filter is used for a grid-modulated stage, condenser input is desirable. A single-section choke input filter does not furnish sufficient ripple suppression, except for a c.w. amplifier or a *push-pull* modulator stage.

Class B Modulator Voltage Regulation. Power supply voltage regulation of class B modulators is of great importance, because the plate current varies appreciably with the amount of speech input. Choke input, utilizing preferably a *swinging-choke* with high no-current inductance rating (25 hy. or more) and low d.c. resistance, in conjunction with mercury vapor rectifiers and a husky filter condenser (at least 4 μfd.), will make a good power supply. If the resting plate current of the modulator tubes is high, as is the case with some of the zero bias class B tubes, a swinging type choke is not essential; however, even so, the choke should have high inductance (10 or 20 hy.).

A comparatively high degree of ripple, as compared to a modulated amplifier power supply, can be tolerated in a power supply feeding a push-pull audio or modulator stage, because a good percentage of the hum is cancelled out in the coupling transformer, if the modulator tubes are well matched.

Power Supplies

ANY device which incorporates vacuum tubes requires a power supply for the filament and plate circuits of the tube or tubes. The filaments of the tubes must be heated in order to produce a source of electrons within the vacuum tubes; direct-current voltages are needed for the other electrodes in order to obtain detection, amplification, and oscillation.

Rectification

Either a.c. or d.c. voltage may be used for filament power supply in most applications; however, the a.c. power supply is the more economical and can be used with most tubes without introduction of hum in the output of the vacuum tube device. The plate potential must be secured from a d.c. source, such as from batteries or a rectified and filtered a.c. power supply.

First the a.c. must be converted into a unidirectional current; this is accomplished by means of vacuum tube *rectifiers,* of either the *full-* or *half-wave* type.

Half-Wave Rectifiers. A half-wave rectifier passes one half of the wave of each cycle of the alternating current and blocks the other half. The output current is of a *pulsating* nature, which can be smoothed into pure, direct current by means of *filter* circuits. Half-wave rectifiers produce a pulsating current which has zero output during one half of each a.c. cycle; this makes it difficult to filter the output properly into d.c. and also to secure good voltage regulation for varying loads.

Full-Wave Rectifiers. A full-wave rectifier consists of a pair of half-wave rectifiers working on opposite halves of the cycle, connected in such a manner that each half of the rectified a.c. wave is combined in the output as shown in Figure 1. This pulsating unidirectional current can be filtered to any desired degree, depending upon the particular application for which the power supply is designed.

A full-wave rectifier consists of two plates and a filament, either in a single glass or metal envelope for low-voltage rectification or in the form of two separate tubes, each having a single plate and filament for high-voltage rectification. The plates are connected across the high-voltage a.c. power transformer winding, as shown in Figure 2. The power transformer is for the purpose of transforming the 110-volt a.c. line supply to the desired secondary a.c. voltages for filament and plate supplies. The transformer delivers alternating current to the two plates of the rectifier tube; one of these plates is positive at any instant during which the other is negative. The center point of the high-voltage transformer winding is usually grounded and is, therefore, at zero voltage, thereby constituting the *negative B connection.*

While one plate of the rectifier tube is conducting, the other is inoperative, and vice versa. The output voltages from the rectifier tubes are connected together through a common rectifier filament circuit, and thus the plates alternately supply pulsating current to the output (load) circuit. The rectifier tube filaments are always positive in polarity with respect to the output in this type circuit.

The output current pulsates 120 times per second for a full-wave rectifier connected to a 60-cycle a.c. line supply, and the output from the rectifier must connect to a *filter,* which will smooth the pulsations into direct current. Filters are designed to select or reject alternating currents; those most commonly used in a.c. power supplies are of the *low-pass* type. This means that pulsating currents which have a frequency below the cutoff frequency of the filter will pass through the filter to the load. Direct current can be considered as alternating current of zero frequency; this passes through the low-pass filter. The 120-cycle pulsations are similar to alternating current in characteristic, so that the filter must be designed to have a *cutoff* at a frequency *lower than 120 cycles* (for a 60 cycle a.c. supply).

Filter Circuits

A low-pass filter consists of combinations of inductance and capacitance. An inductance or *choke coil* offers an impedance to any change

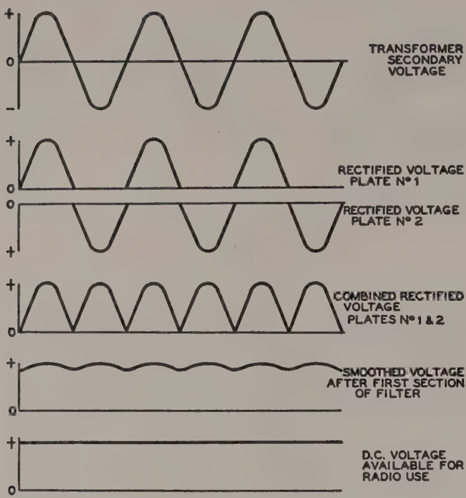


Figure 1.
FULL-WAVE RECTIFICATION.
Showing effects of rectification and filtering of an alternating current. A full-wave rectifier is shown in Figure 2.

in the current that flows through it. A high-inductance choke coil offers a relatively high impedance to the flow of pulsating current, with the result that the *a.c. component* or *ripple* passes from the rectifier tube through the load only with the greatest of difficulty. A capacitance has exactly the opposite action to that of an inductance. It offers a low impedance path to the flow of alternating or pulsating current, but presents practically infinite resistance to the flow of direct current. Inductance coils are usually connected in series with the rectifier outputs, while condensers are connected across the positive and negative leads of the circuit. A simple filter circuit is shown in Figure 3.

Electricity always follows the path of least resistance or impedance. The direct current will travel through the choke and back to the ground (negative B) connection through the *external load,* which normally consists of the plate circuits of vacuum tubes. The a.c. component, or ripple, tends to be impeded by the choke and short-circuited by the condensers across the filter, which offer a lower reactance to the pulsating voltage than that offered by the load. The *load impedance* across the output of most filter systems is generally high, usually from 5,000 to 10,000 ohms. This load resistance can be calculated by dividing the output voltage by the total load current; this value is necessary in making calculations for low-pass resonant types of filter circuits.

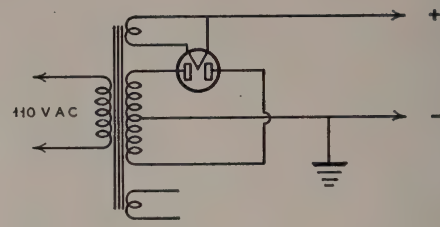


Figure 2.
STANDARD FULL-WAVE SINGLE-PHASE RECTIFIER CIRCUIT.

Resonant Type Filters. In Figure 4, condenser C_1 tunes the choke coil inductance to series resonance at the ripple frequency. Series resonance provides a very low impedance to the resonant frequency limited only by the actual resistance of the choke coil (since the reactance of both the condenser C_1 and the choke coil cancel each other).

The filter circuit in Figure 4 accomplishes the same purpose as a large shunt condenser at the ripple frequency, but is not effective in short-circuiting the higher harmonics in the output of the rectifier system. Additional low-pass filter circuits are needed to remove these harmonic components, which are of great enough magnitude to produce objectionable high-pitched hum in the vacuum tube amplifier circuits.

A typical *low-pass* filter is diagrammed in Figure 5. The combination of C_1, C_2, and L should give a cutoff frequency below that of the rectified output pulsation frequency.

This type of filter is very effective, yet uncritical because the circuit can be designed with *any* cutoff frequency, as long as the attenuation or rejection at the 120-cycle-and-higher harmonic frequencies is great. This type of filter is sometimes called a *"brute force" filter,* because large values of inductance and capacitance are normally used without much attention being paid to the actual cutoff frequency. Inductance values of 10 to 30 henrys are used for filter chokes, and shunt capacities of from 2 to 16 microfarads commonly are used for C_1 and C_2 in Figure 5.

A *resonant trap circuit,* such as shown in Figure 6, is sometimes used to increase the impedance of the choke L at some particular frequency, such as 120 cycles per second.

Parallel resonance of C_2 and L provides a very high impedance at the resonant frequency. The condenser C tends to by-pass the higher ripple harmonics that get through the trap circuit. This type of filter is often used in conjunction with an additional section of filter of the type shown in Figure 3.

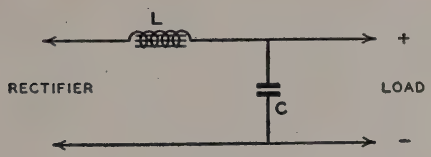


Figure 3.
SIMPLE SINGLE-SECTION CHOKE-INPUT FILTER.
With commonly used values of L and C, the percentage ripple will be between 3 and 10 per cent, depending upon the load resistance. This type of filter is often used to feed a push-pull modulator stage (in which much of the plate voltage ripple cancels out) and telegraphy amplifiers in which slight modulation of the carrier can be tolerated.

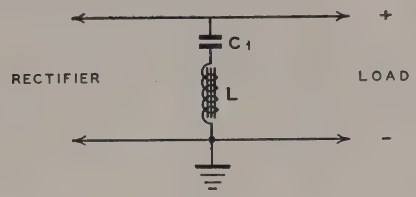


Figure 4.
SERIES RESONANT FILTER CIRCUIT.
If the ripple voltage is high, a high a.c. component will appear across each reactance. With high values of L and correspondingly low values of C, the a.c. voltage across each of these components may exceed the d.c. supply voltage.

The single-section, low-pass filter in Figure 5 is often combined with an additional choke coil as shown in Figure 7. The additional choke coil L_1 is an aid in filtering and also provides better voltage regulation for varying d.c. loads, such as presented by a class B audio amplifier.

A two-section, low-pass filter with condenser input is shown in Figure 8. In some cases, additional sections of choke coils and condensers are added for the purpose of obtaining very pure direct current.

Resistors may be used in place of inductances in circuits where the load current is of low value, or where the applied d.c. voltage must be reduced to some desired value.

The ripple in the output of a filter circuit can be measured with an oscilloscope or by means of the simple circuit in Figure 9. A high-voltage condenser C_3, having a capacity of from $\frac{1}{4}$ to 1 μfd., and a high-resistance copper-oxide a.c. voltmeter provides a method of measuring the actual *ripple voltage.*

The voltmeter should be plugged into the measuring jack after the power supply and external load circuit are in *normal operating condition,* and the meter should be removed from the shorting type jack before turning off the power supply or removing the load. The charging current through condenser C_3 would soon burn out the meter if it were left in the circuit at all times.

Rectifier and Filter Circuit Considerations

The shunt condensers in a filter system serve a dual purpose. They provide: (1) a low impedance path for ripple, (2) an energy-storing system for maintaining constant voltage output from the power supply. The condensers are charged when the peak voltage is applied across them from the output of the rectifier; during the time in which the rectifier output decreases to zero, the filter condensers supply output current to the load. This action provides a constant output voltage.

R.M.S. and Peak Values. In an a.c. circuit, the maximum peak voltage or current is $\sqrt{2}$ or 1.41 times that indicated by the a.c. meters in the circuit. The meters read the *root-mean-square* (r.m.s.) values, which are the peak values divided by 1.41 for a sine wave.

If a potential of 1,000 r.m.s. volts is obtained from a high-voltage secondary winding of a transformer, there will be 1,410-volts peak potential from the rectifier plate to ground. The rectifier tube has this voltage impressed on it, either positively when the current flows or *"inverse"* when the current is blocked on the other half-cycle. The *inverse peak voltage* which the tube will stand safely is used as a rating for rectifier tubes. At higher voltages the tube is liable to arc back, thereby destroying it. The relations between peak inverse voltage, total transformer voltage and filter output voltage depend upon the characteristics

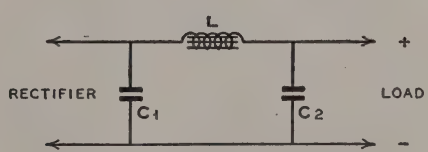


Figure 5.
SINGLE-SECTION CONDENSER INPUT OR π-TYPE FILTER.
This filter is also known as a low-pass or "brute force" filter.

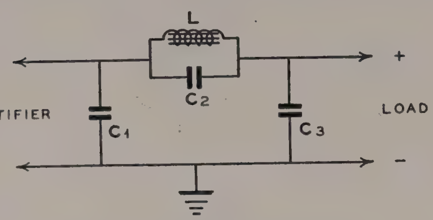


Figure 6.
TUNED FILTER CIRCUIT.

The condensers C_1 and C_3 are the usual values which would be used with a "brute force" filter circuit. The value of C_2 is adjusted to resonate the choke L to the main ripple frequency. This type of filter has very great attenuation to the main ripple frequency, but its attenuation to the ripple frequency harmonics is less than a filter of the "brute force" type. Hence it is advisable to follow a filter of this type with a section of "brute force" filter; very excellent filtering is thus attained.

of the filter and rectifier circuits (whether full- or half-wave, bridge, etc.).

Rectifier tubes are also rated in terms of *peak plate current*. The actual direct load current which can be drawn from a given rectifier tube or tubes depends upon the type of filter circuit. A full-wave rectifier with condenser input may be called upon to deliver a peak current several times the direct load current.

In a filter with choke input, the peak current is not much greater than the load current if the inductance of the choke is fairly high (assuming full wave rectification).

A full-wave rectifier with two rectifier elements requires a transformer which delivers twice as much a.c. voltage as would be the case with a half-wave rectifier or bridge rectifier.

Bridge Rectification. The bridge rectifier is a type of full-wave circuit in which four rectifier elements or tubes are operated from a single high-voltage winding on the power transformer.

While twice as much output voltage can be obtained from a bridge rectifier as from a center-tapped circuit, the permissible output current is only one-half as great for a given power transformer. In the bridge circuit, four rectifiers and three filament heating transformer windings are needed, as against two rectifiers and one filament winding in the center-tapped full-wave circuit. In a bridge rectifier circuit, the inverse peak voltage impressed on any one rectifier tube is halved, which means that tubes of lower peak voltage rating can be used for a given voltage output.

The output voltage across the filter circuit depends upon the design of the filter, resistance of rectifier power transformer, and load resistance. A low-resistance rectifier, such as the mercury-vapor type 83 or 866, has very low voltage drop in comparison with most *high-vacuum* (not mercury-filled) rectifiers. The filter circuit with *condenser input,* i.e., a condenser across the rectifier output, will deliver a higher d.c. voltage than one with *choke input,* but with a sacrifice both in voltage regulation and the amount of available load current.

The d.c. voltage across the load circuit of a condenser-input filter may be as high as 1.4 times the a.c. input voltage (r.m.s.) across one of the rectifier tubes if the input condenser capacity is large and the current drain small. Low values of load resistance (heavy current drain) will cause this type of power supply to have a d.c. voltage output as low or even lower than the a.c. input to the rectifier. The maximum permissible load current in this same circuit is less for a given transformer-secondary wire size and rectifier tube peak current rating than would be the case for a choke-input filter.

A choke-input filter will reduce the d.c. voltage to a value of 0.9 the a.c., r.m.s. value, but the output voltage with choke input is fairly

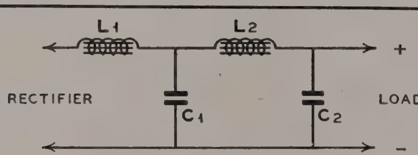


Figure 7.

LOW-PASS "BRUTE FORCE" FILTER WITH INPUT CHOKE.

Adding an input choke to the "brute force" filter improves both regulation and filtering at a sacrifice in output voltage.

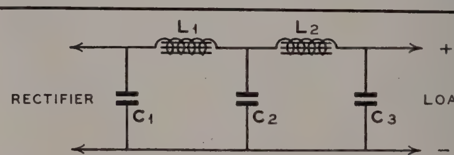


Figure 8.
2-SECTION LOW-PASS FILTER FOR USE WHERE VERY PURE D.C. IS REQUIRED.

This type of power supply filter is widely used in conjunction with low level speech amplifier stages.

constant over a wide range of load resistances, and the allowable load current is greater than with condenser input for a given rectifier and power transformer.

Filter Choke Coils. Filter inductors often consist of a coil of wire wound on a laminated iron or steel core. The size of wire is determined by the amount of direct current which is to flow through the choke coil. This direct current magnetizes the core and reduces the inductance of the choke coil; therefore, filter choke coils of the "smoothing" type are built with an air gap, a small fraction of an inch in the iron core, for the purpose of preventing saturation when maximum d.c. flows through the coil winding.

This "air gap" is usually in the form of a piece of fiber inserted between the ends of the laminations. The air gap reduces the initial inductance of the choke coil, but keeps it at a higher value under maximum load conditions. The coil must have a great many more turns for the same initial inductance when an air gap is used.

As mentioned previously, choke input tends to keep the output voltage of the filter at approximately 0.9 of the r.m.s. voltage impressed upon the filter from the rectifiers. However, this effect does not take place until the load current exceeds a certain minimum value. In other words, as the load current is decreased, at a certain critical point the output voltage begins to soar. This point is determined by the inductance of the input choke. If it has high inductance, the current can be reduced to a very low value before the output voltage begins to rise. Under these conditions, a low-drain bleeder resistor will keep the current in excess of the critical point, and the voltage will not soar even if the external load is removed.

For this purpose, chokes are made with little or no air gap in order to give them more inductance at low values of current. Their filtering effectiveness at maximum current is impaired somewhat, because they saturate easily, but their high inductance at low values of current permits use of a smaller bleeder to keep the current in excess of the critical value. Such chokes are called *swinging chokes* because, while they have high initial inductance, the inductance rapidly falls to a comparatively low value as the current through the choke is increased.

The d.c. resistance of any filter choke should be as low as possible in conjunction with the desired value of inductance. Small filter chokes, such as those used in radio receivers, usually have an inductance of from 6 to 15 henrys, and a d.c. resistance of from 200 to 400 ohms. A high d.c. resistance will reduce the output voltage, due to the voltage drop across each choke coil. Large filter choke coils for radio transmitters and class B amplifiers usually have less than 100 ohms d.c. resistance.

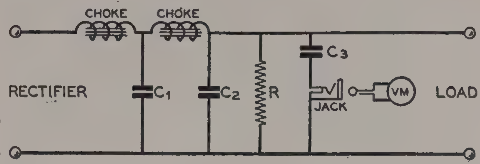


Figure 9.

CIRCUIT FOR MEASURING A.C. RIPPLE.

The meter should not be inserted in the circuit until after the voltage is turned on; otherwise the charging current surge may blow the meter. The jack must be of the closed-circuit type. C_3 must be rated at considerably more than the plate voltage to provide a safety factor.

Filter Condensers. There are two types of filter condensers: (1) paper dielectric type, (2) electrolytic type.

Paper condensers consist of two strips of metal foil separated by several layers of waxed paper. Some types of paper condensers are wax-impregnated; others, especially the high-voltage types, are oil-impregnated. High voltage filter condensers which are oil-impregnated will withstand a greater peak voltage than those impregnated with wax, but they are more expensive to manufacture. Condensers are rated both for *flash* test and normal operating voltages; the latter is the important rating and is the maximum voltage which the condenser should be required to withstand in service.

The condenser across the rectifier circuit in a condenser-input filter should have a working voltage rating equal to at *least* 1.41 times the r.m.s. voltage output of the rectifier. The remaining condensers may be rated more nearly in accordance with the d.c. voltage.

Electrolytic condensers are of two types: (1) wet, (2) dry. The wet electrolytic condenser consists of two aluminum electrodes immersed in a solution called an *electrolyte.* A very thin film of oxide is formed on the surface of one electrode, called the *anode.* This acts as the dielectric. The electrolytic condenser must be correctly connected in the circuit so that the anode always is at positive potential with respect to the electrolyte, the latter actually serving as the other electrode (plate) of the condenser. A reversal of the polarity for any length of time will ruin the condenser.

The dry type of electrolytic condenser uses

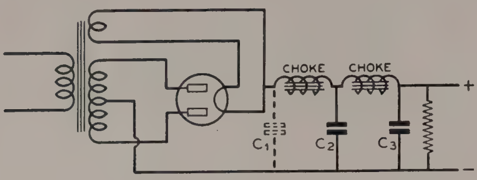


Figure 10.
STANDARD 2-SECTION FILTER.
When C_1 is connected in the circuit, the filter is termed "condenser input." If C_1 is omitted, the filter is called "choke input."

an electrolyte in the form of paste. The dielectric in both kinds of electrolytic condensers is not perfect; these condensers have a much higher direct current leakage than the paper type. The leakage current is greater in the wet electrolytic than in the dry types, but the former are self-healing and are not permanently damaged by moderate voltage overloads.

The high capacitance of electrolytic condensers results from the thinness of the film which is formed on the plates. The maximum voltage that can be safely impressed across the average electrolytic filter condenser is between 450 and 600 volts; the working voltage is usually rated at 450. When electrolytic condensers are used in filter circuits of high-voltage supplies, the condensers should be connected in series. The positive terminal of one condenser must connect to the negative terminal of the other, in the same manner as dry batteries are connected in series.

It is not necessary to connect shunt resistors across each electrolytic condenser section as it is with paper capacitors connected in series, because electrolytic capacitors have fairly low internal d.c. resistance as compared to paper condensers. Also, if there is any variation in resistance, it is that electrolytic unit in the poorest condition which will have the highest leakage current, and therefore the voltage across this condenser will be lower than that across one of the series connected units in better condition and having higher internal resistance. Thus we see that equalizing resistors are not only unnecessary across series connected electrolytic condensers but are actually undesirable. This assumes, of course, similar capacitors by the same manufacturer and of the same capacity and voltage rating. It is *not advisable* to connect in series electrolytic condensers of different make or ratings.

There is very little economy in using electrolytic condensers in series in circuits where more than two of these condensers would be required to prevent voltage breakdown.

Wet electrolytic capacitors housed in an aluminum can ordinarily use the can as the negative electrode, or contact to the electrolyte (the electrolyte being the true electrode). Wet electrolytic condensers should always be mounted in a vertical position. To allow escape of gas generated as a result of electrolysis, a small vent is provided.

Electrolytic condensers can be greatly reduced in size by use of etched aluminum foil for the anode. This greatly increases the surface area, and the dielectric film covering it, but raises the power factor slightly. For this reason, ultra-midget electrolytic condensers should not be used at full rated d.c. voltage when a high a.c. component is present, such as would be the case for the input condenser in a condenser-input filter.

When a dry (paste electrolyte) electrolytic condenser is subjected to over voltage and the leakage current is increased substantially, the condenser may be considered as no longer fit for service, as heating caused by the rupture will aggravate the condition. As previously mentioned, mildly ruptured *wet* electrolytic condensers will heal if normal voltage is applied to them for a time.

Bleeder Resistors. A heavy-duty resistor should be connected across the output of a filter in order to draw some load current at all times. This resistor avoids soaring of the voltage at no load when swinging choke input is used, and also provides a means for discharging the filter condensers when no external vacuum-tube circuit load is connected to the filter. This *bleeder* resistor should normally draw approximately 10 per cent of the full load current.

The power dissipated in the bleeder resistor can be calculated by dividing the square of the d.c. voltage by the resistance. This power is dissipated in the form of heat, and, if the resistor is not in a well-ventilated position, the wattage rating should be higher than the actual

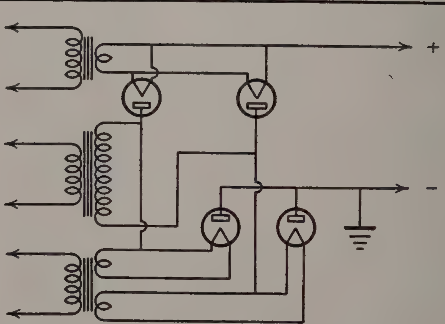


Figure 11.
BRIDGE RECTIFIER CIRCUIT.

wattage being dissipated. High voltage, high capacity filter condensers can hold a dangerous charge if not bled off, and wire-wound resistors occasionally open up without warning. Hence it is wise to place carbon resistors in series across the regular wire-wound bleeder as explained in Chapter 11 under *Safety Precautions.*

When purchasing a bleeder resistor, be sure that the resistor will stand not only the required wattage, but also the *voltage.* Some resistors have a voltage limitation which makes it impossible to force sufficient current through them to result in rated wattage dissipation. This type of resistor usually is provided with slider taps, and is designed for voltage divider service. An untapped, non-adjustable resistor is preferable as a high voltage bleeder, and is less expensive. Several small resistors may be used in series, if desired, in order to obtain the required wattage and voltage rating.

Glow-Discharge Voltage Regulators. Three very useful tubes for stabilizing the voltage on receivers, electron coupled oscillators in exciters, frequency meters, and other devices requiring a constant source of voltage of between 100 and 300 volts are the VR-75-30, VR-105-30, and VR-150-30 glow-discharge type voltage-regulator tubes. These tubes are similar except for their voltage ratings. The remarks following apply generally to all, though the examples apply specifically to the VR-150. All three tubes have the same current rating.

The VR-105 and VR-75 are useful for stabilizing the voltage on the oscillator section of 6J8, 6K8 and similar mixer tubes, for use in the cathode of the feedback tube in a 2A3 type voltage regulated power supply, and many other applications. The VR-150 is suited where higher voltage is desirable.

Two VR type tubes may be connected in series to provide exactly 210, 255, or 300 volts when more than 150 volts is required.

A VR type tube may be used to stabilize the voltage across a variable load or the voltage across a constant load fed from a varying source of voltage. Thus can be seen their many possible applications and wide range of usefulness.

A device requiring, say, only 50 volts can be stabilized against *supply voltage* variations by means of a VR-105 simply by putting a suitable resistor in series with the regulated voltage and the load, dropping the voltage from 105 to 50 volts. However, it should be borne in mind that under these conditions the device will *not* be regulated for *varying load;* in other words, if the *load resistance* varies, the voltage across the load will vary, even though the regulated voltage remains at 105 volts.

To maintain constant voltage across a *varying load resistance* there must be *no* series resistance between the regulator tube and the load. This means that the device must be operated exactly at one of the voltages obtainable by seriesing two or more similar or different VR tubes.

A VR-150 may be considered as a stubborn variable resistor having a range of from 30,000 to 5000 ohms and so intent upon maintaining a fixed voltage of 150 volts across its terminals that when connected across a voltage source having *very poor regulation* it will instantly vary its own resistance within the limits of 5000 and 30,000 ohms in an attempt to maintain the same 150 volt drop across its terminals when the supply voltage is varied. The theory upon which a VR tube operates is covered under the subject of gaseous conduction in the chapter on *Vacuum Tube Theory,* and will not be discussed here.

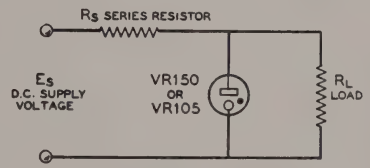


Figure 12.

STANDARD CIRCUIT FOR GLOW-DISCHARGE REGULATOR.

The regulator tube will maintain the voltage across its terminals constant to within 1 or 2 volts for moderate variations in R_L or E_S.

It is paradoxical that in order to do a good job of regulating, the regulator tube must be fed from a voltage source having poor regulation (high series resistance). The reason for this presently will become apparent.

If a high resistance is connected across the VR tube, it will not impair its ability to maintain a fixed voltage drop. However, if the load is made too low, a variable 5000 to 30,000 ohm shunt resistance (the VR-150) will not exert sufficient effect upon the resulting resistance to provide constant voltage except over a *very limited* change in supply voltage or load resistance. The tube will supply maximum regulation, or regulate the largest load, when the source of supply voltage has high internal or high series resistance, because a variation in the effective internal resistance of the VR tube will then have more controlling effect upon the load shunted across it.

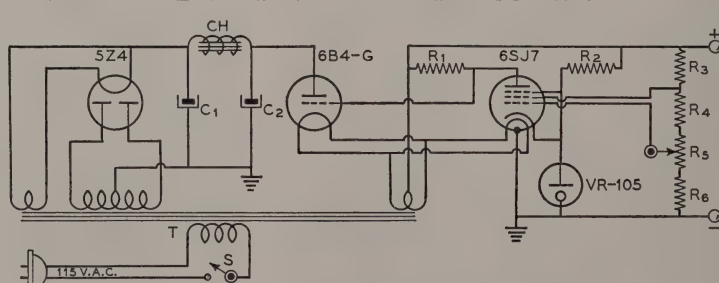
Figure 13.
VOLTAGE REGULATED POWER SUPPLY.
This power pack is capable of putting out from 175 to 300 volts at an average current of 60 ma. with very good stability with respect to both load and supply voltage variations.

In order to provide greatest range of regulation, a VR tube (or two in series) should be used with a series resistor (to effect a poorly regulated voltage source) of such a value that it will permit the VR tube to draw from 15 to 20 ma. under normal or average conditions of supply voltage and load impedance. For maximum control range, the series resistance should be not less than approximately 20,000 ohms, which will necessitate a source of voltage considerably in excess of 150 volts. However, where the supply voltage is limited, good control over a *limited range* can be obtained with as little as 3000 ohms series resistance. If it takes less than 3000 ohms series resistance to make the VR tube draw 15 to 20 ma. when the VR tube is connected to the load, then the supply voltage is not high enough for proper operation.

Should the current through a VR-150, VR-105, or VR-75 be allowed to exceed 30 ma., the life of the tube will be shortened. If the current falls below 5 ma., operation will become unstable. Therefore, the tube must operate within this range, and within the two extremes will maintain the voltage within 1.5 per cent. It takes a voltage excess of at least 10 or 15 per cent to "start" a VR type regulator; and to insure positive starting each time, the voltage supply should preferably exceed the regulated output voltage rating by about 20 per cent or more. This usually is automatically taken care of by the fact that if sufficient series resistance for good regulation is employed, the voltage impressed across the VR tube before the VR tube ionizes and starts passing current is quite a bit higher than the starting voltage of the tube.

When a VR tube is to be used to regulate the voltage applied to a circuit drawing *less than 15 ma.* normal or average current, the simplest method of adjusting the series resistance is to remove the load and vary the series resistor until the VR tube draws exactly 30 ma. Then connect the load, and that is all there is to it. This method is particularly recommended when the load is a heater type vacuum tube, which may not draw current for several seconds after the power supply is turned on. Under these conditions, the current through the VR tube will never greatly exceed 30 ma. even when it is running unloaded (while the heater tube is warming up

Figure 14.
SCHEMATIC OF THE VOLTAGE REGULATED SUPPLY.



C_1—8-μfd. 600-volt electrolytic
C_2—8-μfd. 450-volt electrolytic
R_1—500,000 ohms, ½ watt
R_2—75,000 ohms, 1 watt
R_3—10,000 ohms, 1 watt
R_4—20,000 ohms, 1 watt
R_5—15,000-ohm potentiometer
R_6—10,000 ohms, 1 watt
S—A.c. line switch
T—700 c.t. 120 ma., 5 v. 4 a., 6.3 v. 4.7 a.
CH—10-hy. 110-ma. choke

Figure 15.
VOLTAGE REGULATED GRID BIAS SUPPLY.

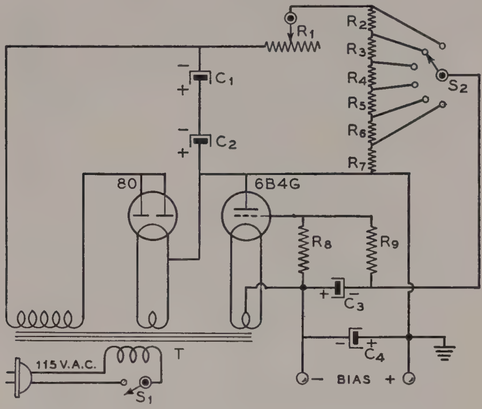


Figure 16.
SCHEMATIC OF THE REGULATED BIAS PACK.

C_1, C_2, C_3, C_4—4-μfd. 450-volt electrolytics
R_1—50,000-ohm potentiometer
R_2, R_3, R_4, R_5, R_6, R_7—50,000 ohms, ½ watt
R_8—100,000 ohms, ½ watt
R_9—10,000 ohms, ½ watt
S_1—A.c. line switch
S_2—Voltage selector switch, s.p. 6-position
T—480 c.t. 40 ma., 5 v. 2 a., 6.3 v. 2 a.

and the power supply rectifier has already reached operating temperature).

Figure 12 illustrates the standard glow discharge regulator tube circuit. The tube will maintain the voltage across R_L constant to within 1 or 2 volts for moderate variations in R_L or E_S.

Voltage Regulated Power Supplies. When it is desired to stabilize the potential across a circuit drawing more than a few milliamperes, it is advisable to use a voltage regulated power supply of the type shown in Figures 13 and 14 rather than glow discharge type tubes. The power pack illustrated will deliver up to 300 volts of well-regulated voltage, the output voltage holding within 1 volt for variations in line voltage or load resistance of 25 per cent.

The maximum current that may be drawn from the supply without detrimentally affecting the regulation is determined by the desired output voltage, the latter being adjustable by variation of R_3. At 200 volts the output voltage is constant up to 100 ma., the maximum current which the 6B4-G and power transformer will stand. At 300 volts, the maximum usable output voltage, the useful range is from 0 to 50 ma. At the latter voltage the regulator begins to lose control when more than 50 ma. is drawn from the supply.

The system works by virtue of the fact that the 6B4-G acts as a variable series resistance or losser, and is controlled by a regulator tube much in the manner of a.v.c. circuits or inverse feedback as used in radio receivers and a.f. amplifiers. The 6SJ7 amplifier controls the bias on the 6B4-G, which in turn controls the resistance of the 6B4-G, which in turn controls the output voltage, which in turn controls the plate current of the 6SJ7, thus completing the cycle of regulation. It is readily apparent that under these conditions any change in the output voltage will tend to "resist itself," much as the a.v.c. system of a receiver resists any change in signal strength delivered to the detector.

Because it is necessary that there always be a moderate voltage drop through the 6B4-G in order for it to have proper control, the rest of the power supply is designed to deliver as much output voltage as possible considering the r.m.s. voltage of the b.c.l. type power transformer. This calls for a low resistance full-wave rectifier, a high capacity input condenser, and a low d.c. resistance filter choke. A 5Z4 rectifier is used in place of an 83 or other mercury vapor tube to avoid possible "hash" in any nearby receiver. This tube has lower resistance than an 80 or 5Z3 and in addition, since it is a heater type, plate voltage will not be applied to the regulator tubes until they are up to operating temperature.

Voltage-Regulated Bias Pack. The type of voltage-regulated power supply discussed in the previous paragraphs is not suited for use as a bias pack. Since the direction of current flow in a bias power supply is opposite from

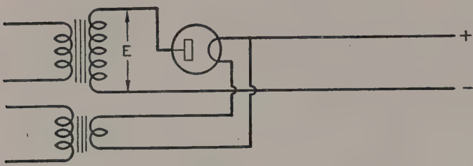


Figure 17.
SINGLE-PHASE HALF-WAVE RECTIFIER.
The output from this type rectifier is not easily filtered except where very little current is drawn (assuming 25 to 60 cycle supply).

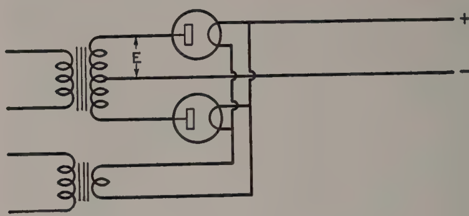


Figure 18.
SINGLE-PHASE FULL-WAVE RECTIFIER.

that of a regular power supply, a special type of pack must be used for bias service. A suitable pack for use in a regulated bias circuit is shown in Figures 15 and 16. In this type of power supply, the regulator tube (6B4-G, 2A3, or 6A3) acts as a variable *bleeder* resistor which automatically adjusts its resistance to a value such that the grid current flowing through it will develop a constant value of voltage across the output terminals of the pack.

Inspection of the circuit diagram, Figure 16, will show that the circuit consists of a half-wave power supply (to obtain greater voltage from the b.c.l.-type power transformer), a pair of electrolytic condensers in series as the filter, and a tapped voltage divider feeding the grid of the 6B4-G regulator tube. The tap switch, S_2, provides a rough voltage adjustment from about 100 to 600 volts while the rheostat, R_1, allows a fine voltage adjustment to be made. The maximum grid current which may be run through the pack is determined by the plate dissipation of the 6B4-G. The permissible grid current varies from about 100 ma. in the vicinity of 100 volts of bias down to about 25 ma. in the 600-volt region.

Rectifier Circuits

The three types of rectifier circuits for single-phase a.c. line supply consist of a half-wave rectifier, as shown in Figure 17, a full-wave rectifier as shown in Figure 18, and a bridge rectifier circuit as shown in Figure 19.

Three-phase circuits can be connected for half-wave rectification, as shown in Figure 20, or for full-wave rectification as shown in Figure 21.

The most popular circuits are those shown in Figures 18 and 19. The maximum transformer voltage of the high-voltage secondary, d.c. output voltage for choke-input filter, and maximum direct load current are shown in the accompanying table in terms of rectifier tube peak ratings. These peak ratings are listed in a separate table for a few commonly used rectifier tubes.

As an example, suppose type 866-A rectifier tubes are used as in Figure 18: The maximum transformer voltage E across each side of the center tap is 0.35 times 10,000 or 3,500 volts. The d.c. voltage at the input to the filter (choke input) is 3,500 times 0.9 or 3,150 volts. The maximum advisable d.c. output current is 0.66 times the peak plate current of 1.0 ampere or 660 milliamperes.

These are the maximum voltages and currents which can be used without exceeding the ratings of the rectifier tubes. The actual d.c. voltage at the output of the filter will depend upon the d.c. resistance of the filter, and can be found by subtracting the IR drop across the filter chokes from the value of 0.9 times the transformer voltage E. This does not take into consideration the voltage drop in the power transformer and rectifier tubes. The voltage drop across a mercury vapor rectifier tube is always between 10 and 15 volts. However, the voltage drop across high-vacuum rectifier tubes can be many times greater.

The power supply circuits illustrated in Figures 22 to 25 represent commonly-used connections for power transformers. The values of d.c. output voltage are indicated in

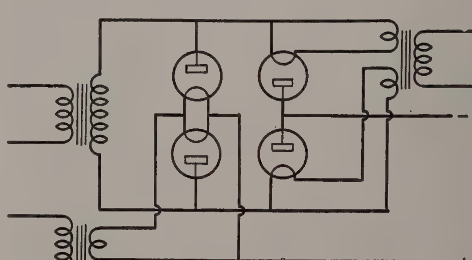


Figure 19.
BRIDGE CIRCUIT.

TUBE TYPE	PEAK INV. VOLTAGE	PEAK PLATE CURRENT
82	1550	.345
83	1550	.675
866 Jr.	5000	.500
816	5000	.500
866A/866	10000	1.000
249-B	10000	1.500
KY-21	11000	3.000 (grid)
RX-21	11000	3.000
872	7500	5.000
872-A	10000	5.000
869	20000	5.000

each case for a load current of 100 ma. The transformer secondary potential is 1,100 volts. The interesting figures in connection with each circuit are those of the primary current.

The circuit in Figure 25 should never be used unless the load current is very low. Manufacturers generally rate their transformers in terms of secondary r.m.s. voltage and the maximum d.c. load current which can be taken *from a choke input filter* circuit such as shown in Figure 22. In order to prevent overload of the power transformer in Figure 25, the load current must be reduced to less than one third of the value which can be drawn from the circuit in Figure 22. The load which can be drawn from the circuit in Figure 24 without overload to the power transformer is approximately 50 per cent of that for the circuit in Figure 22. The permissible direct load current in Figure 23 would only be two-thirds as much as for Figure 22, for a given transformer size.

Mercury Vapor Rectifier Tubes. When new or long-unused high-voltage rectifier tubes of the mercury vapor type are first placed in service, the filaments should be operated at normal temperature for approximately 20 minutes before plate voltage is applied, in order to remove all traces of mercury from the cathode. After this preliminary operation, plate voltage can be applied within 20 to 30 seconds of the time the filaments are turned on each time the power supply is used. If plate voltage is applied before the filament is brought to full temperature, active material may be knocked off the oxide-coated filament, and the life of the tube will be greatly shortened.

Small r.f. chokes must sometimes be connected in series with the plate leads of mercury vapor rectifier tubes in order to prevent the generation of radio-frequency hash. These r.f. chokes must have sufficiently heavy wire to carry the load current, and enough inductance to attenuate the r.f. parasitic noise current from flowing into the filter supply leads and radiating into nearby receivers.

Small resistors or small iron-core choke coils should be connected in series with each plate lead of a mercury-vapor rectifier tube when used in circuits such as those shown in Figure 26.

These resistors tend to prevent one plate from carrying the major portion of the current. *High-vacuum* type rectifiers which are connected in parallel do not require these resistors or chokes.

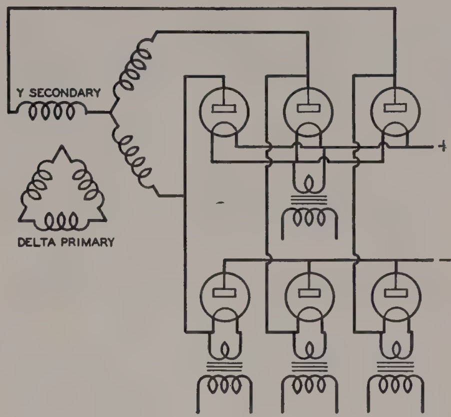


Figure 21.

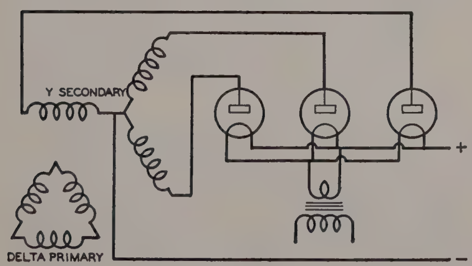
FULL-WAVE 3-PHASE RECTIFIER.

Even with no filter the output of this rectifier will have a high percentage of d.c. A simple filter will suppress the small amount of ripple, because of the high ripple frequency (6 times supply frequency).

Figure 20.

HALF-WAVE 3-PHASE RECTIFIER.

The output of this rectifier has a d.c. component and a higher ripple frequency (3 times supply frequency), and therefore is easier to filter than a single-phase full-wave rectifier.



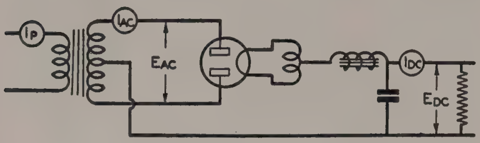


Figure 22.

FULL-WAVE RECTIFIER—CHOKE INPUT.

E_{DC} — 435 v. E_{AC} — 1100 v.
I_{DC} — 100 ma. I_{AC} — 71 ma.
I_{PRI} — 0.6 a.

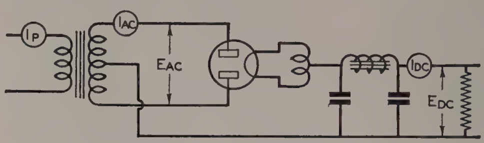


Figure 23.

FULL WAVE RECTIFIER—CONDENSER INPUT.

E_{DC} — 675 v. E_{AC} — 1100 v.
I_{DC} — 100 ma. I_{AC} — 103 ma.
I_{PRI} — 0.9 a.

Bias Voltage Power Supplies. Power packs to supply negative grid voltage for radio or audio amplifiers differ from plate supplies mainly in that the positive and negative connections are reversed; the positive terminal of a C-bias supply is connected to ground. The filter chokes are usually connected in series with the hot (ungrounded) lead, which in this case is the *negative lead.*

The bias voltage supply for a linear r.f. amplifier or class B audio amplifier must have a very low resistance bleeder. The bleeder should be chosen so that the normal bleeder current is at least 8 times the *peak* grid current of the class B modulator or linear r.f. amplifier. If this condition is not met, the bias pack will act somewhat as a grid leak and the bias on the tubes will rise excessively under modulation.

High μ tubes require so little bias and draw so much grid current for class B operation (either r.f. or a.f.) that battery bias is ordinarily employed. It is inadvisable to use a bias pack for this purpose unless the required bias voltage is more than 90 volts.

When a pack having a tapped bleeder is used for this type of service, the tap on the bleeder should be by-passed for voice frequencies, even though the pack already has a large filter condenser across the outside terminals of the bleeder.

High efficiency grid modulation also requires a low resistance source of bias, though the bias voltage required is usually several times as great as for a class B stage using tubes of similar power. For this reason, bias for this type of amplifier is more commonly obtained from a regulated bias pack rather than from a conventional pack utilizing a very low resistance bleeder in order to comply with the requirement of low resistance in the bias supply. Such a regulated bias pack is described earlier in this chapter. Another is described in Chapter 8 under *High Efficiency Grid Modulation.*

Bias Pack Considerations. It should be borne in mind that when a power supply is used "inverted" in order to provide bias to a stage drawing grid current, the grid current flows in the *same direction as the bleeder current.* This means that the grid current does not flow through the power pack as when a pack is used to supply plate voltage, but rather through the bleeder. The transformer and chokes in the bias pack actually have less work to do when the biased stage is drawing

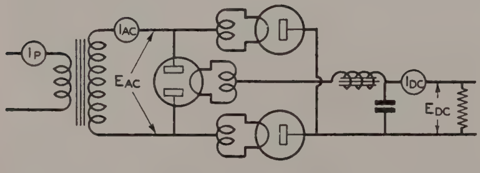


Figure 24.

BRIDGE RECTIFIER—CHOKE INPUT.

E_{DC} — 860 v. E_{AC} — 1100 v.
I_{DC} — 100 ma. I_{AC} — 96 ma.
I_{PRI} — 1.1 a.

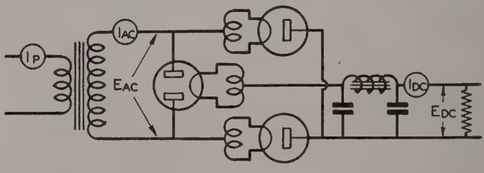


Figure 25.

BRIDGE RECTIFIER—CONDENSER INPUT.

E_{DC} — 1200 v. E_{AC} — 1100 v.
I_{DC} — 100 ma. I_{AC} — 148 ma.
I_{PRI} — 1.65 a.

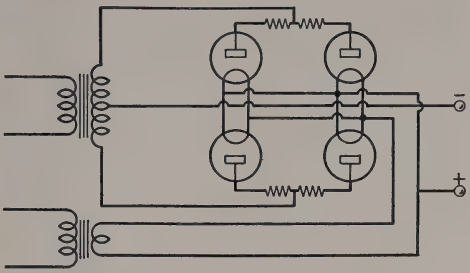


Figure 26.

PARALLEL OPERATION OF MERCURY-VAPOR RECTIFIERS.

Small, center-tapped resistors or iron core chokes are used to make the current divide evenly. If not used, one rectifier of each parallel pair tends to take the whole load. 100 ohms or 1 hy., center tapped, is satisfactory for each equalizer.

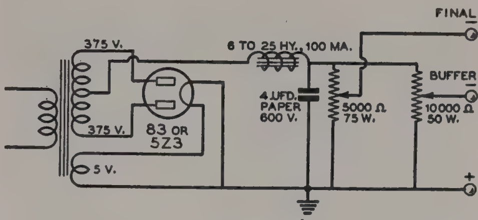


Figure 27.

BIAS PACK FOR C.W. OR PLATE MODULATED TRANSMITTER.

This pack will deliver up to 250 volts of protective bias to the various stages of a high-power 'phone or c.w. transmitter. A large safety factor in the filter condenser is provided to permit using the full output of the pack for bias as would be the case with crystal keying of high power using medium-μ tubes and running heavy grid current. The power transformer should be of about 75 ma. rating. This type of bias pack does not have good regulation, and should not be used with class B linear or class B audio stages; such applications require a very low resistance bleeder. To bias a grid modulated amplifier, another section of filter should be added.

grid current, because the greater the grid current flowing through the bleeder the greater the voltage drop across it and the less current the bias pack supplies to the bleeder. In fact, if the grid current is great enough and the bleeder resistor high enough, the voltage developed across the bleeder will be greater than the maximum voltage which the power pack can deliver, and hence the power pack will be delivering no current to the bleeder. Under these conditions, it is quite possible for the voltage to exceed the voltage rating of the bias pack filter condensers.

Bear in mind that the bleeder always acts as a grid leak when grid current is flowing, and while the effect can be minimized by making the resistance quite low, all grid current *must flow through the bleeder,* as it cannot flow back through the bias pack.

Class C amplifiers, both c.w. and plate modulated, require high grid current and considerably more than cutoff bias, the bias sometimes being as high as 4 or 5 times cutoff. To protect the tubes against excitation failure, it is desirable that fixed bias sufficient to limit the plate current to a safe value be used. This is normally the amount of bias that would be used on the same tubes at the same plate voltage in a class B modulator. It is best practice to obtain only this amount of bias from a bias pack, the additional required amount being obtained from a variable grid leak which is adjusted for correct bias and grid current while the stage is running under normal conditions.

This condition is such that the voltage divider tap on the bias pack will be delivering only a portion of the full bias pack voltage when the biased stage is inoperative. Then, when grid current flows to the biased stage, there is no danger of the voltage rising to dangerously high values across the filter condensers in the bias pack.

A bias power supply for providing "protective bias" to the r.f. stages of a medium-power radio transmitter is shown in Figure 27.

Two bleeder resistors with slider adjustments provide any desired value of negative grid bias for the r.f. amplifiers. The location

FIGURE NO.	TRANSFORMER VOLTS MAX. "E"	D.C. OUTPUT VOLTS AT INPUT TO FILTER	D.C. OUTPUT CURRENT IN AMPERES
18	.35 x Inv. Pk. Vtg.	.9 x E	.66 x Pk. Plate
19	.7 x Inv. Pk. Vtg.	.9 x E	.66 x Pk. Plate
20	.43 x Inv. Pk. Vtg.	1.12 x E	.83 x Pk. Plate
21	.43 x Inv. Pk. Vtg.	2.25 x E	1.0 x Pk. Plate

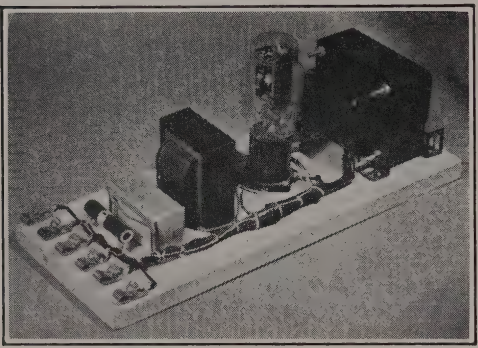


Figure 28.
TYPICAL 300-350 VOLT POWER SUPPLY.

This type of power pack ordinarily utilizes a power transformer having from 325 to 350 volts each side of c.t. with integral filament windings, and a brute force or pi-type filter consisting of a single choke and dual 8-μfd. electrolytic condenser. The rectifier is usually an 80 or 5Z3. Such packs commonly deliver from 50 to 150 ma., depending upon the ratings of the transformer and choke, and are most commonly used with receivers, a.f. amplifiers and drivers, and low power exciter stages. They are also used as bias packs, in which case a very low resistance bleeder is used, adjustable taps being provided so that the bleeder can be used as a voltage divider.

of the slider on the resistors should be determined experimentally with the amplifier in operation, since the direct grid current of the r.f. amplifier itself will affect the voltage across the bias supply taps. The circuit illustrated is practically free from reaction between buffer and final amplifier bias.

Transmitter Power Input Control

In the interests of interference reduction, one should run only sufficient power input to a radio transmitter to maintain satisfactory communication. The power input to the final r.f. amplifier of a c.w. transmitter can be controlled over a very wide range by means of an *autotransformer,* connected as in Figure 29.

The a.c. voltage can be varied from a few volts up to 130 volts, by means of a relatively small autotransformer. This a.c. voltage should be applied only to the high-voltage power transformer which supplies plate power to the final r.f. amplifier.

Convenient adjustment of input to a phone transmitter other than of the plate modulated type is a more difficult problem. Input to a plate-modulated transmitter can be varied the same as for a c.w. transmitter without danger of overmodulating the reduced input, if the primary voltage for the plate transformer that feeds the modulators is fed from the same tap on the autotransformer as the plate transformer for the final amplifier. This assumes the modulators are of the "zero bias" type. If one power supply is used for both, the problem is further simplified.

Reducing the power of a grid-modulated final amplifier is more of a problem. The best method for reducing power is to reduce the r.f. excitation *and* audio gain together, without disturbing the bias or plate voltage or antenna coupling adjustment.

Those using linear r.f. amplifiers can either incorporate a switching arrangement for throwing the antenna over to the low-level modulated stage and thus reduce power about 10 db, or else merely reduce excitation to the linear amplifier *without* disturbing the a.f. gain control.

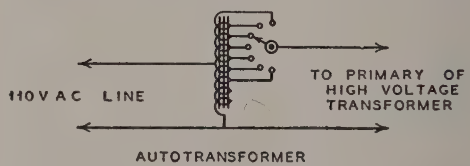


Figure 29.
AUTOTRANSFORMER VOLTAGE CONTROL.

Overload Protection

To protect the tubes in a medium or high power final amplifier in the event of excitation failure, any one of several courses is satisfactory. If very high μ tubes of the "zero bias" type commonly used as modulators are used in the class C amplifier, no protection is necessary if the tubes are run within their voltage rating. However, such tubes ordinarily require somewhat more excitation than an equivalent medium high μ tube. Thus, an 811 requires more excitation than an 812, nearly twice as much when plate modulated.

Safety bias from a bias pack or batteries invariably is used on the various stages in a c.w. transmitter which is oscillator keyed. However, when the final amplifier is keyed, or for telephony, such bias is not required from the standpoint of keying, but simply to protect the tubes from excessive dissipation in the event of accidental excitation failure. Other means of protecting the tubes against such possibility are as follows:

Overload relays, which can be adjusted to trip and stay open at any desired amount of plate current, can be used to open the primary of the plate transformer. However, such relays are rather expensive as compared to

simple relays of the s.p.s.t. type, and usually cost as much as a bias pack.

A small instrument fuse (*Littelfuse*) of appropriate current rating can be placed in the *center tap* lead to the amplifier stage (the grid leak going to ground or B minus side of the fuse). These fuses cost but 10 cents, and while rated at only 250 volts, will work satisfactorily at high plate voltage when placed in the center tap lead and not in the B plus or B minus lead. The current rating should be such that the fuse does not blow immediately when the plate current is excessive, but does blow before the tubes become hot enough to be damaged. The correct value to use can be determined by blowing one or two fuses experimentally by detuning the amplifier or otherwise making it draw a momentarily large plate current.

These methods of protection (overload relay or fuse) are satisfactory only when the amplifier tube draws *considerably more current at zero bias than it does under normal operating conditions.* This generally will apply with tubes having a μ of 30 or less. It is obvious that if the tube draws about the same plate current at zero bias and no excitation as it does under normal conditions, the dissipation can be excessive without the fuse or overload relay being actuated. By looking up the characteristic curves on a tube it is possible to ascertain if it draws appreciably more plate current at zero bias and no excitation than it does under normal operating conditions.

Perhaps the best type of protection is obtained by means of an inexpensive s.p.s.t. relay which is actuated by the grid current to the final stage. The relay winding should be such that the contacts close at about half normal grid current, and the winding should be capable of handling somewhat more than normal grid current without damage. Suitable relays are those sold by several manufacturers for less than $1.75, and a suitable winding can be had for almost any grid current from 25 to 200 ma. by the choice of 6, 12, and 24 volt d.c. windings available.

The relay contacts may be used to short out a cathode bias resistor. In this way, the amplifier has cathode bias until excitation is applied, then the relay closes and shorts out the cathode bias. Used in this manner, the relay contacts have little work to do. Just sufficient cathode bias should be used to keep the plate dissipation from exceeding the rated maximum when excitation fails. Normally about 750 ohms will be about right for one tube and 400 ohms for two tubes.

If the relay contacts are sufficiently heavy, they may be used to break the primary of the high voltage plate transformer feeding the final amplifier. Because of the inductive "kick", rather heavy contacts will be required in this service when running high power.

Because the cathode bias resistor is shorted out when normal plate current is being drawn, the resistor need not have a high dissipation

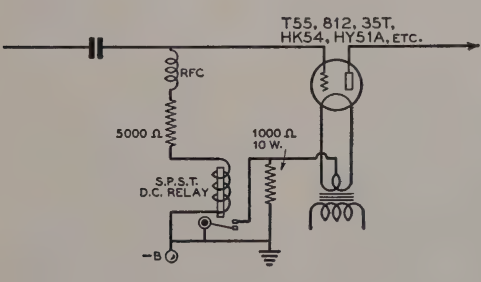


Figure 30.
IMPROVED CATHODE BIAS SYSTEM.

At the cost of a small inexpensive relay, the protection of cathode bias may be obtained with none of the disadvantages of the latter. When excitation is applied, the cathode bias resistor is shorted out. This circuit shows a typical application. The relay should close at about half normal grid current and be capable of standing the maximum grid current.

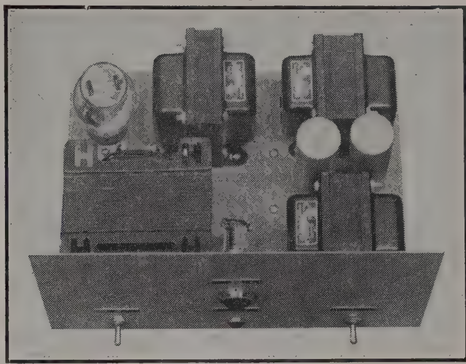


Figure 31.
TYPICAL 500-600 VOLT POWER SUPPLY.

Power supplies such as this are commonly used to feed low power r.f. stages, modulators, etc. The power transformer is generally rated at from 600 to 750 volts each side of c.t. at from 150 to 250 ma. and has no filament windings. An 83 or 5Z3 and swinging choke input filter having 600-volt oil-filled paper condensers are ordinarily used. Round can condensers of this type are usually less expensive than equivalent ones in square cans.

rating. A 10-watt resistor will be large enough for a 250-watt rig. A typical circuit using this arrangement is shown in Figure 30.

Typical Power Supplies

Several photographs and diagrams of power supplies are shown throughout this chapter to illustrate the more common methods of construction. Figures 13 and 14 show a voltage regulated power supply of the vacuum tube regulated type with a 6B4-G as the regulator tube. Figures 15 and 16, respectively, show a bias pack of the stabilized type, also using a 6B4-G as the regulator tube. Figure 28 illustrates a simple 300-350 volt supply of the type commonly used to supply filament and plate voltages to a receiver, v.f.o., or to the low-level stages in a transmitter. Figures 31, 32, 33, and 34 illustrate methods of construction of medium power plate supplies for 600, 1500 and 600, and 1500 volts of the types usually built for supplying plate voltage to the intermediate and final amplifiers of amateur transmitters.

Transformer Design

A common problem in radio and allied work is to determine how a transformer can be built to supply certain power requirements for a particular application, or how to calculate the windings needed to fit a certain transformer core which is already on hand. These problems can be solved by a small amount of calculation.

The most important factor in determining the size of any transformer is the amount of core material available. The electrical rating, as well as the physical size, is determined almost entirely by the size of the core. The core material is also important. The present practice is to use high-grade silicon-steel sheet. It will be assumed that this type of material is to be employed in all construction herein described. Soft sheet-iron or stovepipe iron is sometimes substituted, but transformers made from such materials will have about 50 to 60 per cent of the power rating, pound for pound of core, as those made from silicon-steel.

The Core. The core size determines the performance of a transformer because the entire energy circulating in the transformer (except small amounts of energy dissipated in resistance losses in the primary) must be transformed from electrical energy in the primary winding to magnetic energy in the core, and reconverted into electrical energy in the secondary. The amount of core material determines quite definitely the power that any transformer will handle.

Transformer cores are often designed so that if the losses per cubic inch of core material are determined, these losses can be used as a basis for calculating the rating of the transformer. These losses exist in watts, and are divided between the eddy current loss and the hysteresis loss. The eddy current loss is the loss due to the lines of force moving across the core, just as if it were a conductor, and setting up currents in it.

Induced currents of this type are very undesirable and they are merely wasted in heat-

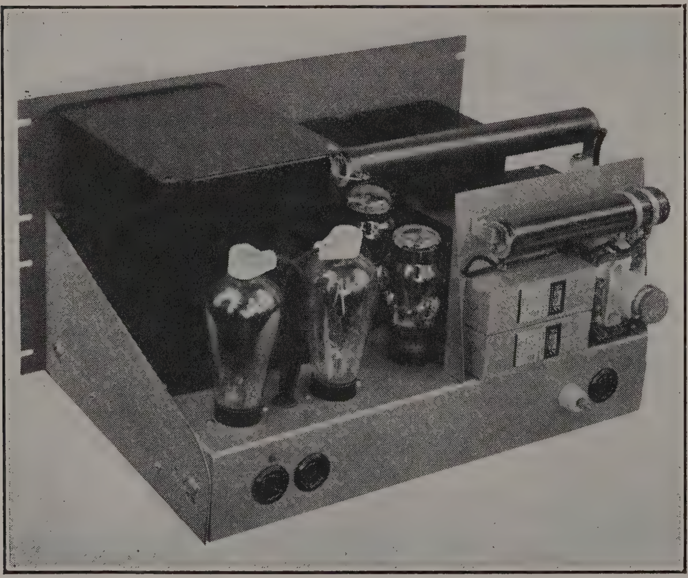


Figure 32.

POWER SUPPLY FOR RELAY RACK MOUNTING.

Illustrating well-designed dual power pack for relay rack mounting. Observe the fuse, bleeders, and the feedthrough terminal for the high voltage connection. This unit delivers 1500 and 600 volts. Note that the heaviest components are mounted towards the front panel to minimize the strain on panel and chassis.

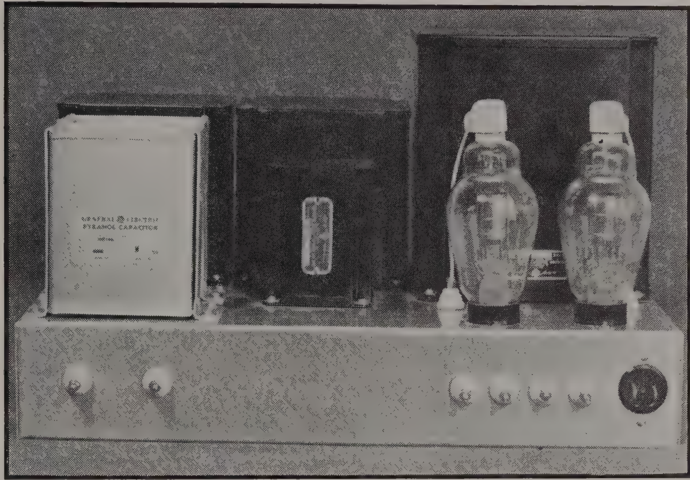


Figure 33.
GENERAL PURPOSE 1500-VOLT POWER SUPPLY.

Taps are provided on the secondary of the power transformer so that the output voltage of the power supply may be reduced to 1250 or 1000 volts should this be desirable. The permissible current drain from the supply at either of the three voltages is 300 ma.

ing the core, which then tends to heat the windings, increase the resistance of the coils, and reduce the overall power handling ability of the transformer. To reduce such losses, transformer cores are made of thin sheets, usually about no. 29 gauge. These sheets are insulated from each other by a coat of thin varnish, shellac or japan, or by the iron-oxide scale which forms on the sheets during the manufacturing process and which forms a good insulator between sheets.

Hysteresis. The magnetic flux in the core lags behind the magnetizing force that produces it, which is, of course, the primary supply. Because all transformers operate on alternating current, the core is subjected to continuous magnetizing and demagnetizing force, due to the alternating effect of the a.c. field. This *hysteresis* (meaning "to lag") heats the iron, due to molecular friction caused by the iron molecules re-orienting themselves as the direction of the magnetizing flux changes.

Saturation. The higher the field strength, the greater the heat produced. A condition can be reached where a further increase in magnetizing flux does not produce a corresponding increase in the flux density. This is called "saturation," and is a condition which would cause considerable heat in a core. In practice, it has been found that all core material must be operated with the magnetic flux well below the limit of saturation.

Core Losses. All core losses manifest themselves as heat, and these losses are the determining factor in transformer rating. They are spoken of as "total core loss," generally used as a single figure, and for common use a core loss of from .75 watt to 2.5 watts per pound of core material can be assumed for 60 cycles. The lower figure is for the better grades of thin sheet, while the higher loss is for heavier grades.

About 1 watt per pound is a very satisfactory rating for common grades of material. This rating is also dependent on the manner in which the transformer is built and mounted, and on the ease with which the heat is radiated from the core. Transformers with higher losses may be used for intermittent service.

The transformer core loss can be assumed to be from 5 to 10 per cent of the total rating for small transformers. Thus, if the core loss is known, the rating of the transformer can be easily determined. If the figure of 1 watt per pound is assumed, the problem is further simplified. To determine the rating of the transformer, weigh the core. If, for example, the core weighs 10 pounds, the transformer will handle from 100 to 200 watts. Such a transformer core can be assumed to have about 150 watts nominal rating.

If the weighing of the core is inconvenient, the weight can be calculated from the cubic content or volume. Sheet-steel core laminations weigh approximately one-fourth pound per cubic inch.

Transformer cores are generally made in two types: shell, and core. The shell-type has a center leg which accommodates the windings, and this is twice the cross-sectional areas of the side legs. The core-type is made from strips built up into a hollow-like affair of uniform cross section. For the shell-type core, the area is taken as the square section of the center leg, in this case $2\frac{1}{4} \times 4\frac{1}{2}$ inches and in the core-type, this area is taken as the section of one leg, and is also $2\frac{1}{4} \times 4\frac{1}{2}$ inches, or an actual core area in both cases of 10.1 square inches, which is large enough for a comparatively large transformer.

Copper Wire Table

Gauge No. B. & S.	Diam. in Mils[1]	Circular Mil Area	Turns per Linear Inch[2]				Turns per Square Inch[2]			Feet per Lb.		Ohms per 1000 ft. 25° C.	Correct Capacity at 1500 C.M. per Amp.[3]	Diam. in mm.
			Enamel	S.S.C.	D.S.C. or S.C.C.	D.C.C.	S.C.C.	Enamel	D.C.C.	Bare	D.C.C.			
1	289.3	82690	—	—	—	—	—	—	—	3.947	—	.1264	55.7	7.348
2	257.6	66370	—	—	—	—	—	—	—	4.977	—	.1593	44.1	6.544
3	229.4	52640	—	—	—	—	—	—	—	6.276	—	.2009	35.0	5.827
4	204.3	41740	—	—	—	—	—	—	—	7.914	—	.2533	27.7	5.189
5	181.9	33100	—	—	—	—	—	—	—	9.980	—	.3195	22.0	4.621
6	162.0	26250	—	—	—	—	—	—	—	12.58	—	.4028	17.5	4.115
7	144.3	20820	—	—	—	—	—	—	—	15.87	—	.5080	13.8	3.665
8	128.5	16510	7.6	—	7.4	7.1	—	—	—	20.01	19.6	.6405	11.0	3.264
9	114.4	13090	8.6	—	8.2	7.8	—	—	—	25.23	24.6	.8077	8.7	2.906
10	101.9	10380	9.6	—	9.3	8.9	87.5	84.8	80.0	31.82	30.9	1.018	6.9	2.588
11	90.74	8234	10.7	—	10.3	9.8	110	105	97.5	40.12	38.8	1.284	5.5	2.305
12	80.81	6530	12.0	—	11.5	10.9	136	131	121	50.59	48.9	1.619	4.4	2.053
13	71.96	5178	13.5	—	12.8	12.0	170	162	150	63.80	61.5	2.042	3.5	1.828
14	64.08	4107	15.0	—	14.2	13.8	211	198	183	80.44	77.3	2.575	2.7	1.628
15	57.07	3257	16.8	—	15.8	14.7	262	250	223	101.4	97.3	3.247	2.2	1.450
16	50.82	2583	18.9	18.9	17.9	16.4	321	306	271	127.9	119	4.094	1.7	1.291
17	45.26	2048	21.2	21.2	19.9	18.1	397	372	329	161.3	150	5.163	1.3	1.150
18	40.30	1624	23.6	23.6	22.0	19.8	493	454	399	203.4	188	6.510	1.1	1.024
19	35.89	1288	26.4	26.4	24.4	21.8	592	553	479	256.5	237	8.210	.86	.9116
20	31.96	1022	29.4	29.4	27.0	23.8	775	725	625	323.4	298	10.35	.68	.8118
21	28.46	810.1	33.1	32.7	29.8	26.0	940	895	754	407.8	370	13.05	.54	.7230
22	25.35	642.4	37.0	36.5	34.1	30.0	1150	1070	910	514.2	461	16.46	.43	.6438
23	22.57	509.5	41.3	40.6	37.6	31.6	1400	1300	1080	648.4	584	20.76	.34	.5733
24	20.10	404.0	46.3	45.3	41.5	35.6	1700	1570	1260	817.7	745	26.17	.27	.5106
25	17.90	320.4	51.7	50.4	45.6	38.6	2060	1910	1510	1031	903	33.00	.21	.4547
26	15.94	254.1	58.0	55.6	50.2	41.8	2500	2300	1750	1300	1118	41.62	.17	.4049
27	14.20	201.5	64.9	61.5	55.0	45.0	3030	2780	2020	1639	1422	52.48	.13	.3606
28	12.64	159.8	72.7	68.6	60.2	48.5	3670	3350	2310	2067	1759	66.17	.11	.3211
29	11.26	126.7	81.6	74.8	65.4	51.8	4300	3900	2700	2607	2207	83.44	.084	.2859
30	10.03	100.5	90.5	83.3	71.5	55.5	5040	4660	3020	3287	2534	105.2	.067	.2546
31	8.928	79.70	101.	92.0	77.5	59.2	5920	5280	—	4145	2768	132.7	.053	.2268
32	7.950	63.21	113.	101.	83.6	62.6	7060	6250	—	5227	3137	167.3	.042	.2019
33	7.080	50.13	127.	110.	90.3	66.3	8120	7360	—	6591	4697	211.0	.033	.1798
34	6.305	39.75	143.	120.	97.0	70.0	9600	8310	—	8310	6168	266.0	.026	.1601
35	5.615	31.52	158.	132.	104.	73.5	10900	8700	—	10480	6737	335.0	.021	.1426
36	5.000	25.00	175.	143.	111.	77.0	12200	10700	—	13210	7877	423.0	.017	.1270
37	4.453	19.83	198.	154.	118.	80.3	—	—	—	16660	9309	533.4	.013	.1131
38	3.965	15.72	224.	166.	126.	83.6	—	—	—	21010	10666	672.6	.010	.1007
39	3.531	12.47	248.	181.	133.	86.6	—	—	—	26500	11907	848.1	.008	.0897
40	3.145	9.88	282.	194.	140.	89.7	—	—	—	33410	14222	1069	.006	.0799

[1]A mil is 1/1000 (one thousandth) of an inch.
[2]The figures given are approximate only, since the thickness of the insulation varies with different manufacturers.
[3]The current-carrying capacity at 1000 C.M. per ampere is equal to the circular-mil area (Column 3) divided by 1000.

Table courtesy P. R. Mallory & Co.

Turns Per Volt. To determine the number of turns for a given voltage, apply the following formula:

$$E = \frac{4.44\,N\,B\,A\,T}{10^8}$$

Where E equals the volts of the circuit; N, the cycles of the circuit; B, the number of magnetic lines per square inch of the magnetic circuit; A, the number of square inches of the magnetic circuit; and T, the number of turns.

The proper value for B, for small transformers and for ordinary grades of sheet-iron, such as are now being considered, is 75,000 for 25 cycles and 50,000 for 50 or 60 cycles.

Rewriting the above formula

$$T = \frac{E \times 10^8}{4.44\,N\,B\,A}$$

and since N and B are known

$$T = \frac{10^8}{4.44 \times 60 \times 50{,}000} \times \frac{E}{A}$$

from which

$$T = 7.5 \times \frac{E}{A}$$

That is, for a transformer to be used on a 60-cycle circuit, the proper number of turns for the primary coil is obtained by multiplying the line voltage by 7.5 and dividing this product by the number of square inches cross section of the magnetic circuit.

On a 25-cycle circuit, the 7.5 becomes 12, and on 50 cycles it becomes 9.

Design Example. Assume a transformer core that is to be used on a 115-volt, 60-cycle circuit for supplying power to two rectifier tubes, each of which takes 1,000 volts on the plate. The rectifier is of the full-wave type. The core measures $2\frac{1}{2} \times 4\frac{1}{2}$ inches; hence,

$$T = \frac{7.5 \times 115}{2.25 \times 4.5} = 85$$ (to the nearest turn), and the volts per turn equals

$$\frac{115}{85} = 1.353$$ which is the same for all coils.

Now, the secondary coil must have two windings in series, each to give 1,000 volts, and with a middle tap. The secondary turns will be $\frac{2000}{1.353} = 1478$ with a tap taken out at the 739th turn.

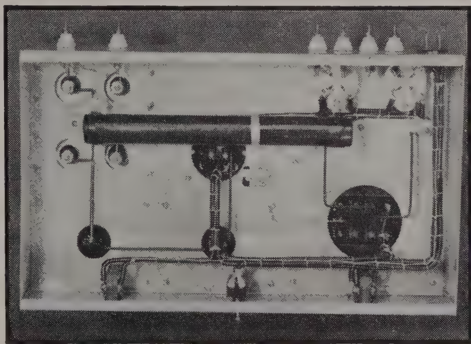


Figure 34.
UNDER-CHASSIS VIEW OF THE 1500-VOLT SUPPLY.

C_1, C_2—5-μfd. 2000-volt oil-filled condensers
R—75,000-ohm 200-watt bleeder resistor
T_1—1760/1460/1250 volts each side of c.t., 300 ma.
T_2—2.5 v. c.t., 10 a.; 5000 volts test
CH_1—5-15 hy., 300-30 ma. swinging choke
CH_2—10-hy. 300-ma. filter choke

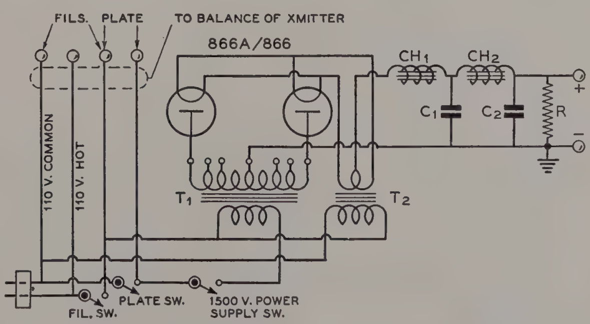


Figure 35.

SCHEMATIC OF THE 1500-VOLT POWER SUPPLY.

Transformer Design Chart

WATTS	Section of Core (inches)	Area of Core (Square inches)	Primary Turns	Primary Wire Size	Turns per Volt	SECONDARY WINDINGS (Turns for Voltages Given)																	
											HIGH-VOLTAGE WINDING												
						2.5 volts	5.0 volts	6.3 volts	7.5 volts	10 volts	250 volts	300 volts	350 volts	400 volts	450 volts	500 volts	600 volts	700 volts	800 volts	900 volts	1000 volts	1250 volts	1500 volts
10	½ x ½	.25	3500	31	32	80	160	205	240	320													
10	½ x ⅝	.31	2800	31	24.2	61	122	147	182	242													
12	½ x ¾	.37	2300	30	20.0	50	100	126	150	200													
12	⅝ x ⅝	.38	2280	30	19.6	48	96	124	147	196													
15	⅝ x ¾	.46	1875	29	16.1	42	84	105	124	161													
22	⅝ x 1	.62	1400	28	12.2	31	61	77	92	122													
20	¾ x ¾	.55	1570	28	13.6	34	68	86	102	136													
25	¾ x 1	.75	1150	27	10.0	25	50	63	75	100	2620	3150	3700	4200	4750	5250							
30	¾ x 1¼	.93	930	26	8.1	21	42	52	62	81	2100	1500	3140	3400	3800	4200							
50	¾ x 1½	1.12	770	24	6.7	17	34	43	50	67	1860	2100	2500	2840	3150	3500	4200	5000					
50	1 x 1	1.0	860	24	7.5	19	38	48	57	75	1950	2400	2700	3150	3600	3900	4700	5500					
60	1 x 1¼	1.25	690	23	6.0	15	30	38	45	60	1600	1900	2200	2500	2800	3150	3800	4400					
65	1 x 1½	1.50	575	23	5.0	13	25	32	38	50	1300	1575	1850	2100	2400	2650	3150	3700					
75	1 x 1¾	1.75	490	22	4.2	11	21	27	31	42	1100	1320	1550	1750	2000	2200	2650	3150	3800	4000	4400		
110	1 x 2	2.0	430	21	3.7	9	18	23	28	37	980	1170	1370	1550	1750	1960	2300	2750	3100	3500	3900		
105	1¼ x 1¼	1.56	550	21	4.8	12	24	31	36	48	1260	1510	1770	2050	2240	2510	3050	3500	4100	4500	5020		
100	1¼ x 1½	1.87	460	21	3.8	9	19	25	29	38	1000	1200	1400	1600	1800	2000	2400	2720	3200	3560	4000		
120	1¼ x 1¾	2.18	400	20	3.5	9	18	21	26	35	920	1100	1315	1470	1650	1840	2200	2560	2940	3300	3700	4620	5500
140	1¼ x 2	2.5	350	19	3.2	8	16	20	24	32	840	1020	1180	1340	1510	1680	2050	2350	2680	3000	3380	4200	5050
125	1½ x 1½	2.25	380	20	3.3	8	16	21	25	33	870	1040	1210	1400	1560	1730	2100	2420	2800	3120	3500	4400	5250
150	1½ x 1¾	2.64	330	18	2.9	7	14	19	22	29	760	910	1130	1220	1360	1530	1840	2100	2450	2750	3050	3800	4650
200	1½ x 2	3.0	290	17	2.42	6	12	15	18	24	630	765	890	1020	1150	1265	1522	1780	2050	2380	2350	3200	3840
300	2 x 2	4.0	215	15	1.87	5	9	12	14	19	490	590	690	780	880	980	1180	1360	1570	1760	1950	2350	2940
400	2 x 2½	5.0	175	14	1.52	4	8	10	12	15	395	470	550	640	710	790	950	1110	1265	1420	1590	1980	2400
500	2 x 3	6.0	145	13	1.26	3	6	8	9	12	330	395	455	530	595	660	790	920	1060	1200	1330	1650	2000

Allowing 1,500 circular mils per ampere, the primary wire should be no. 12. The size of the wire on the plate coils may be no. 22 or 24 for a 400 to 300 ma. rating.

To determine the quantity of iron to pile up for a core, it is well to consider 1 to 1.5 volts per turn as a conservative range. For trial, assume 1.25 volts. Then by transforming the first equation

$A = 7.5 \times \frac{E}{T}$ or, the area required is 7.5 times the volts per turn; in this case, $7.5 \times 1.25 = 9.38$ square inches.

The magnetic cross section must be measured at right angles to the laminations that are enclosed by the coil, the center leg when the core is built up around the coil, and either leg where the core is built up inside the coil, that is, between the arrows in the sketches shown on page 339.

It should be kept in mind that there is a copper or resistance loss in all transformers. This is caused by the passage of the current through the windings, and is commonly spoken of as the "IR" loss. It manifests itself directly as heat and varies as the load is varied; the heavier the load, the more heat is developed.

This heat, as well as other heat losses, must be removed, or the transformer will burn up. Most transformers are so arranged that both the core and windings can radiate heat into the surrounding air and thus cool themselves. Large transformers are mounted in oil for cooling, and also for the purpose of increasing the insulation factor.

In any transformer, the voltage ratio is directly proportional to the turns ratio. This means that if the transformer is to have 110-volts input and 250 turns for the primary, and if the output is to be 1,100 volts, 2,500 turns will be needed. This may be expressed as:

$$\frac{E_p}{E_s} = \frac{T_p}{T_s}$$

It is often more convenient to take the figure obtained for the primary winding and, by dividing by the supply voltage, the number of turns per volt is calculated. This accomplished, the number of turns for any given voltage can be calculated by simple multiplication.

Radio transformers are generally of small size. The matter of power factor can therefore be disregarded, more especially because they work into an almost purely resistive load. In the design of radio transformers, the power factor can be safely assumed as unity, in which case the apparent watts and the actual watts are the same. Admittedly, this is not always

Figure 36.
COMBINED MODULATOR AND POWER SUPPLY.

While ordinarily it is preferable to design a power supply as an integral independent unit, it sometimes is desirable for reasons of limited space to construct a modulator or r.f. unit with its power supply on the same chassis. In this example a modulator and its power supply are mounted on a single chassis, thus making use of every bit of space. Also, external connecting cables are avoided, but special care must be taken to prevent hum being picked up from the power supply by the input stage of the speech amplifier.

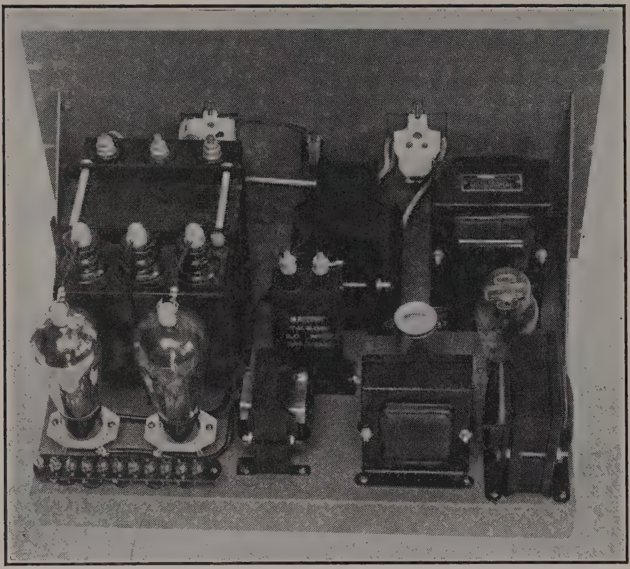


Figure 37.
DUAL POWER SUPPLY.
In a rack-mounted power supply, which is supported from the front panel, it is advisable to place the heavy components close to the front panel to minimize the strain on the panel and chassis. Note the position of the heavy plate transformer. The chassis contains a 350-volt power supply and a 1250-volt power supply. Low voltage power supply components are to the right.

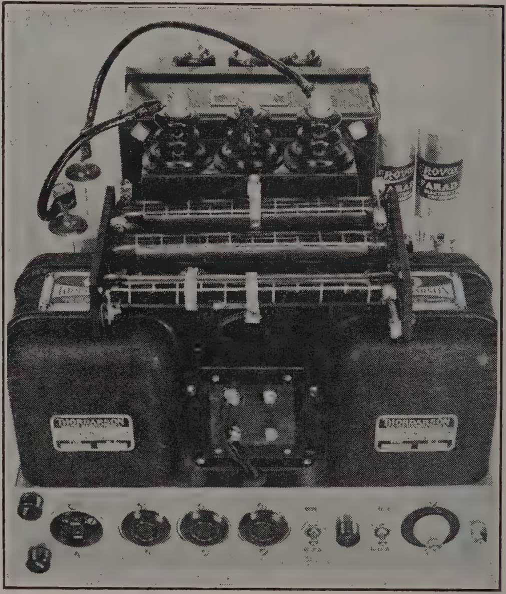
a correct assumption, but it will suffice for common applications.

The size of the wire to be used in any transformer depends upon the amperage to be carried. For a current of 1 ampere as a continuous load, at least 1,000 circular mils per ampere must be allowed. For transformers which have poor ventilation, or continuous heavy load service, or where price is not the first consideration, 1,500 circular mils per ampere is a preferable figure. If, for example, a transformer is rated at 100-watts primary load on 110 volts, the current will be

$$I = \frac{W}{V} = \frac{100}{110} = 0.90 \text{ amperes}$$

and if the assumption is 1,000 circular mils per ampere, it will be found that this will require 1,000 × .90, or 900 circular mils. The wire table on page 334 shows that no. 20 wire for 1,200 mils is entirely satisfactory. If it is desired to use 1,500 circular mils, instead of 1,000, this will require 1,500 × .90 or 1,350 mils, which corresponds to approximately no. 19 wire. The difference seems to be small, yet it is large enough to reduce heating and to improve overall performance. Assume, for tentative design, a 600-volt, 100-ma. high-voltage secondary; a 3-ampere 5-volt secondary; and 2.5-volt 7.5-ampere secondary. Simple calculation will show a 60-watt load on the high-voltage secondary, 15 watts on the 5-volt winding, and 16 watts on the 2.5-volt winding; a total of 91 watts. The core and copper loss

Figure 38.
ILLUSTRATING COMPACT POWER SUPPLY CONSTRUCTION.
This dual power supply illustrates what can be crowded on a chassis when space is at a premium. Note how the heavy bleeders, which also act as voltage dividers, are mounted so as to provide free circulation of air. A pair of 866's serve as rectifier in a conventional circuit for a high voltage supply; an 83 rectifier is used in the low voltage supply.



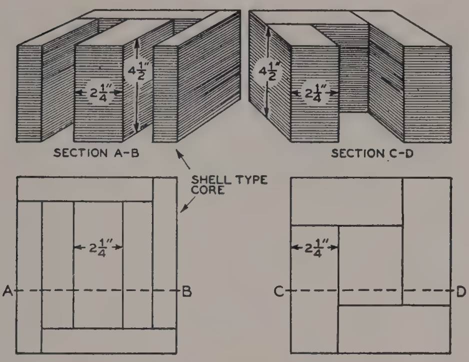


Figure 39.
TYPES OF TRANSFORMER CORES.

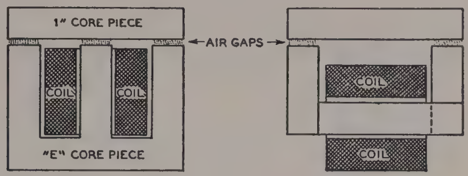


Figure 40.
Two types of choke coil construction. The air gap is approximately 1/32 inch. The gap may be filled with non-magnetic material, such as brass, bakelite, etc.

is 10 watts. The wire sizes for the secondaries will be for 100-ma. current, no. 30 wire; 3 amperes at 5 volts, no. 15 wire; no. 11 wire for the 7.5-ampere secondary.

For high-voltage secondary windings, a small percentage of turns should be added to overcome the resistance of the small wire used, so that the output voltage will be as high as anticipated. The figures given in the table include this percentage which is added to the theoretical ratio and, consequently, the number of turns shown in the table can be accepted as the actual number to be wound on the core of any given transformer.

Insulation. Allowance should always be made for the insulation and size of the windings. Good insulation should be provided between the core and the windings and also between each winding and between turns. Numerous materials are satisfactory for this purpose; varnished paper or cloth, called empire, is satisfactory, although costly. Good bond paper will serve well as an insulating medium for small transformer windings.

Insulation between primary and secondary and to the core must be exceptionally good, as well as the insulation between windings. Thin mica or micanite sheet is very good. Thin fibre, commonly called fish paper, is also a good insulator; bristol board, or strong, thin cardboard may also be used. In all cases, the completed coil should be impregnated with insulating varnish, and either dried in air or baked in an oven. Common varnishes or shellac are unsatisfactory on account of the moisture content of these materials. Air-drying insulating varnish is practical for all-around purposes; baking varnish may be substituted, but the fumes given off are inflammable and often explosive. Care must be exercised in the

CHOKE TABLE FOR TRANSMITTER POWER SUPPLY UNITS

CURRENT M.A.	WIRE SIZE	NO. TURNS	LBS. WIRE	APPROX. CORE (Area)	AIR GAP	WT. CORE
200	No. 27	2000	1.5	1½"x1½"	3/32"	4 lbs.
250	No. 26	2000	1.75	1½"x2"	3/32"	5 lbs.
300	No. 25	2250	2	2"x2"	⅛"	6 lbs.
400	No. 24	2250	3	2"x2½"	⅛"	7 lbs.
500	No. 23	2500	4	2½"x2½"	⅛"	10 lbs.
750	No. 21	3000	6	2½"x3"	⅛"	14 lbs.
1000	No. 20	3000	7.5	3"x3"	⅛"	18 lbs.

NOTES: These are approximately based on high-grade silicon steel cores with total air gaps as given. Air gaps indicated are total of all gaps.

The use of standard "E" and "I" laminations is recommended. If strips are used, and if an ordinary square core is used, the number of turns should be increased about 25%. Choke coils built as per the above table will have an approximate inductance of 10 to 15 henrys. Because considerable differences occur due to winding variations, allowable flux densities of cores, etc., the exact inductance cannot be stated; these chokes will, however, give satisfactory service in radio transmitter power supply systems.

The wire used is based on 1000 circular mils per ampere; this will cause some heating on long runs, and if the chokes are to be used continuously, as in a radiotelephone station in continuous service, it is good practice to use the next size larger choke shown for such loads.

handling of this type of material. Collodion and banana oil lacquer are positively dangerous, and in the event of a short circuit or transformer burn-out, a serious fire may result.

If it is desired to wind a transformer on a given core, it is much better to calculate the actual space required for the windings, then determine whether there is enough available space on the core. If this precaution is not observed, the designer may find that only about half the turns are actually wound on the core, when the space is about three-fourths filled. From 15 to 40 per cent more space than calculated must be allowed. The winding of transformers by hand is a laborious process. Unless the builder is an experienced coil-winder, there is every chance that a sizable portion of the space will be used up by insulation, etc., not sufficient space remaining for the winding. Calculate the cubical space needed for the total number of turns, and allow from 15 to 40 per cent additional space in the core window. Thereby much time and labor will be saved.

Filter Choke Considerations

A choke is a coil of high inductance. It offers an extremely high impedance to alternating current, or to current which is substantially alternating, such as pulsating d.c. delivered at the output of a rectifier.

Choke coils are used in power supplies as part of the complete filter system in order to produce an effectively-pure direct current from the pulsating current source, that is, from the rectifier. The wire size of the choke must be such that the current flowing through it does not cause an appreciable voltage drop due to the ohmic resistance of the choke; at the same time, sufficient inductance must be maintained to provide ample smoothing of the rectified current.

Smoothing Chokes. The function of a smoothing choke is to discriminate as much as possible between the a.c. ripple which is present and the desired d.c. that is to be delivered to the output. Its air gap should be large enough so that the inductance of the choke does not vary materially over the normal range of load current drawn from the power supply, but no larger than necessary to give maximum inductance at full current rating.

Swinging Chokes. In certain radio circuits the power drawn by a vacuum tube amplifier can vary widely. Class B audio amplifiers are good examples of this type of amplifier. The plate current drawn by a class B audio amplifier can vary 1000 per cent or more. It is desirable to keep the d.c. output voltage applied to the plate of the amplifier as constant as possible, and the voltage should be independent of the current drawn from the power supply. The output voltage from a given power supply is always higher with a condenser input filter than with a choke-type input filter. When the input choke is of the *swinging* variety, it means that the inductance of the choke varies widely with the load current drawn from the power supply, due to the fact that high initial inductance is obtained by utilizing a "butt" gap, or none at all as in a transformer core.

Choke Design and Construction

A choke is made up from a silicon-steel core which consists of a number of thin sheets of steel, similar to a transformer core, but wound with only a single winding. The size of the core and the number of turns of wire, together with the air gap which must be provided to prevent the core from saturating, are factors which determine the inductance of a choke. The relative sizes of the core and coil determine the amount of d.c. which can flow through the choke without reducing the inductance to an undesirable low value due to magnetization.

The same core material which is used in ordinary radio power transformers, or from those which are burned out, is satisfactory for all general purposes.

In construction, the choke winding must be insulated from the core with a sufficient quantity of insulating material so that the highest peak voltages which are to be experienced in service will not rupture the insulation.

Transmitter Construction

THE units shown in this chapter are complete transmitters, including modulator and power supply. The complete transmitters are shown for the benefit of those who prefer to construct a whole transmitter from a tried and proven circuit which has been engineered from the standpoint of an integral unit rather than attempt to work out an individual design from the exciter, amplifier, power supply, and modulator units shown elsewhere in this book.

All but one of the complete transmitters shown in this chapter are equipped for radiophone work, because it is a simple matter to omit the modulation equipment if 'phone operation is not desired.

40-WATT TRANSMITTER-EXCITER

The unit illustrated in Figures 1, 2, 4, and 5 is intended to serve as a complete 'phone-c.w. transmitter, or as an r.f. and audio driver for a higher power final amplifier and modulator. The transmitter is designed to provide utmost flexibility, including bandswitching and provision for crystal control from its own crystal oscillator or excitation from a separate variable-frequency oscillator.

R.F. Section

The r.f. section, which is placed at the top of the $17\frac{1}{2}$-inch rack-cabinet, employs a 6L6 as a crystal oscillator followed by another 6L6 as a doubler-quadrupler and an HY-69 amplifier-doubler output stage. All bands from 160 to 10 meters are covered through the use of both stage switching and coil switching.

Bandswitching. For 160-meter operation, a crystal in that band is placed in the crystal socket on the panel and switch S_1 (Figure 3) is thrown to the upper position. With S_1 in this position, the HY-69 is excited directly from the crystal oscillator plate circuit, the doubler-quadrupler stage being cut out of the circuit. Either 160- or 80-meter crystals may be used when 80-meter output is desired, with the HY-69 being operated either as a straight amplifier or doubler.

To reach 40 meters, S_1 is thrown to the lower position, cutting in the second 6L6. An 80-meter crystal is used for this band. The doubler-quadrupler plate circuit is tuned to 20 meters for 20- or 10-meter output and the HY-69 used as a straight amplifier on 20 meters or doubler to 10 meters.

The high capacity plate tank condensers in the two 6L6 stages allow plenty of leeway in winding coils for these stages which will hit two adjacent bands. Although the condensers are somewhat larger than they need to be to cover two bands, their cost is but little greater than that of condensers which will "just cover" the required frequency range. The small additional cost is easily offset by the reduction of coil-winding difficulties. Battery bias is used on the doubler-quadrupler and output stages to permit the use of crystal keying and to allow the doubler-quadrupler to be

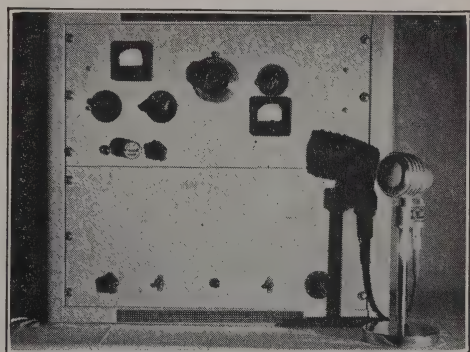


Figure 1.
40-WATT EXCITER-TRANSMITTER.
By itself, this unit forms a complete 40-watt 'phone-c.w. transmitter. With the r.f. amplifier and modulator unit shown later in this chapter it forms a 200-watt transmitter for 'phone or c.w. operation on bands from 10 to 160 meters.

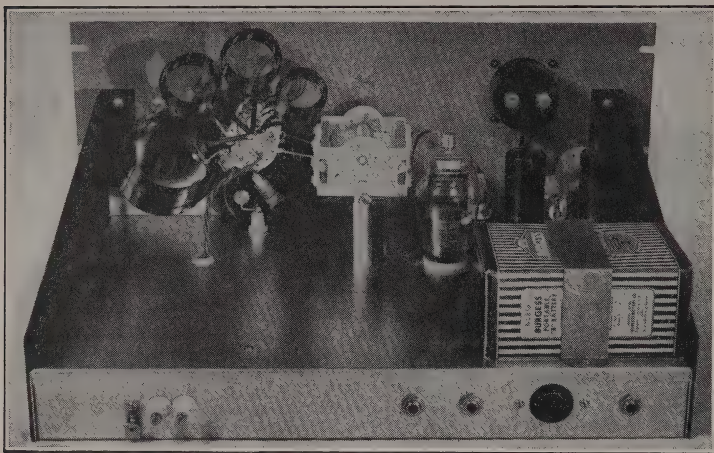


Figure 2.
R.F. SECTION CHASSIS.
The r.f. components of the exciter-transmitter are located on this chassis, which is located at the top of the cabinet shown in Figure 1. The battery at the rear of the chassis supplies fixed bias to the exciter stage, to allow crystal keying.

operated without excitation on the 80- and 160-meter bands. The battery is mounted on the r.f. chassis, and since the current through the battery is small, it may be expected to give long life.

Coil Turret. A manufactured coil-turret assembly is used in the plate circuit of the HY-69 stage. The turret is composed of four coils, separate coils being used for the 10-, 20-, and 40-meter bands, while a single tapped coil is used to cover the 80- and 160-meter bands. One change is required in the coil assembly, as supplied by the manufacturer, to adapt it to use in this transmitter. This change simply involves removing 1 turn from the 2-turn coupling loop on the 10-meter coil, since tests show that the maximum efficiency of transfer to the antenna or following stage is obtained when the coupling coil is reduced to 1 turn. When used as a low power transmitter, the r.f.

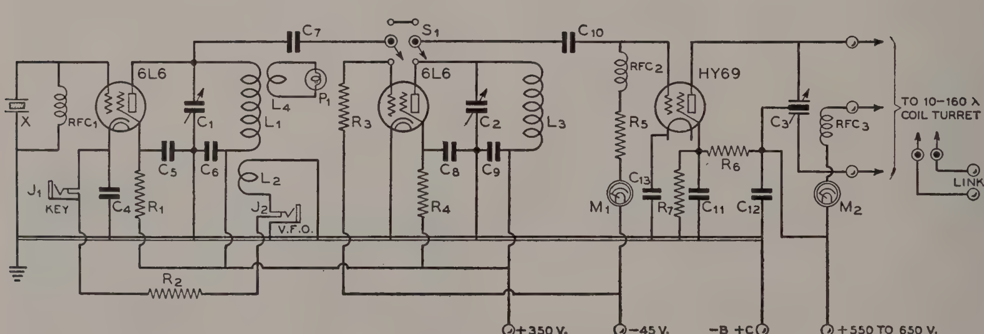


Figure 3.
R.F. SECTION DIAGRAM.

C_1, C_2 — 320-μμfd. midget variable
C_3—265-μμfd. per section, .070″ spacing
C_4, C_5, C_6—.01-μfd. 600-volt tubular
C_7—.001-μfd. mica
C_8, C_9 — .004-μfd. mica
C_{10} — .0005 - μfd. mica
C_{11}—.002-μfd. mica
C_{12}, C_{13}—.004-μfd. mica
R_1 — 15,000 ohms, 10 watts
R_2—600 ohms, 10 watts
R_3—250,000 ohms, 2 watts
R_4—15,000 ohms, 10 watts
R_5—40,000 ohms, 2 watts
R_6—25,000 ohms, 10 watts
R_7—50,000 ohms, 2 watts
RFC_1, RFC_2, RFC_3 —2.5 mhy., 125 ma.
J_1, J_2 — Closed-circuit jack
S_1—D.p.d.t. selector switch, laminated bakelite insulation
M_1—0-15 ma.
M_2—0-150 ma.
L_1—1½ inch long, close-wound with no. 22 d.c.c. on 1″ dia. form
L_2—2 turn link at cold end of L_1
L_3—12 turns no. 18 d.c.c. 1″ dia. and wound to a length of 1½″
X—160- or 80-meter crystal

Figure 4.
BOTTOM VIEW OF R.F. CHASSIS.
The location of the plate coils for the two 6L6 stages is clearly shown in this photograph. It is important that the twisted leads connecting to the r.f. output terminals at the rear of the chassis be kept well separated from the HY 69 grid circuit.

unit should be coupled to the antenna by means of a universal coupler, as shown in Chapter 20.

Metering. The two meters visible on the panel read the plate and grid currents on the output stage. The grid meter serves to show when the doubler-quadrupler stage is tuned to resonance, no other meter being needed for this purpose. When the output stage is operated on the 20- and 10-meter bands, where the excitation is from the doubler-quadrupler, it is helpful to have some indication of the operation of the crystal oscillator stage, however, and the pilot light at the lower left corner of the panel makes a convenient, inexpensive indicator. The 150-ma., 6.3-volt pilot lamp (brown bead) is coupled to the crystal stage plate coil through a 1-turn loop, L_4.

V.F.O. Operation. Means for exciting the transmitter-exciter from a variable frequency oscillator is provided by a 2-turn coil, L_2, around the ground end of the crystal stage plate coil. When an ordinary phone plug is used to terminate the link from the v.f.o., placing the plug in J_2 couples the v.f.o. to L_1 and, at the same time, opens the cathode circuit of the crystal oscillator circuit by breaking the circuit between R_2 and ground. The crystal stage plate tank circuit acts as a tuned grid circuit for the second 6L6 or the HY-69 when a v.f.o. is used.

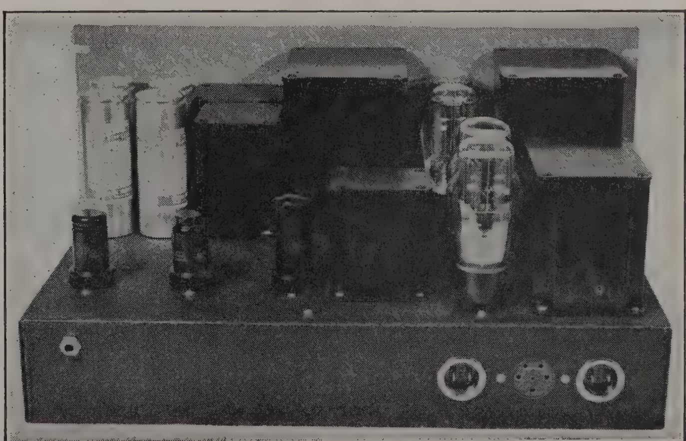
Power Supply and Audio Chassis

The lower chassis in the rack, which, like the r.f. chassis, measures 13 x 17 inches, mounts the audio and power supply section of the transmitter-exciter.

Audio Section. The audio section of the transmitter is intended to serve as a modulator for the HY-69, to form a complete 'phone transmitter, or as a driver for a class B modulator, when the r.f. section is used as an exciter for a medium-power final amplifier. Although the normal output rating for 6A3's is only 10 watts, it is possible to obtain nearly three times this output by driving the grids somewhat and using a low value of plate load. This amount of output is sufficient to fully modulate a plate input of 60 watts to the HY-69. The modulation transformer secondary is merely connected in series with the plate supply to the HY-69, to use the unit as a complete 'phone transmitter.

As the wiring diagram shows, the circuit of the speech amplifier is strictly conventional. The amplifier is designed to give full output

Figure 5.
THE AUDIO AND POWER SUPPLY SECTION.
All of the audio and power supply components are located on this, the lower chassis in the rack. Outlets at the rear of this chassis are provided for the microphone, line voltage, cable to r.f. section and external switch.



with diaphragm-type crystal microphones (−45 to −50 db output level). High level moving-coil (dynamic) microphones will also supply sufficient input to the speech amplifier, if this type is preferred. The 6SJ7 grid resistor, R_1, should be replaced by a line-to-grid transformer if a moving-coil microphone is used. Since the speech amplifier and the power supply are on the same chassis, it will probably be necessary to revolve the input transformer while listening to the output of the amplifier to determine the mounting position which results in minimum hum pickup.

Power Supply. A single transformer rated at 460 volts a.c. each side of the center tap at 325 milliamperes is used in the dual-voltage power supply. To handle the 300 milliamperes of current drawn by the complete transmitter-exciter, two type 83 rectifiers are used. One of the rectifiers operates into a condenser-input filter and delivers 600 volts at 100 milliamperes to the HY-69 stage. The other 83 rectifier delivers voltage to a two-section, choke-input filter and thence to the 6L6 r.f. stages and to the speech amplifier-modulator. Plate voltage for the 6A3 audio output is taken from the junction of the two filter chokes following the latter rectifier.

Filament transformer T_2 supplies all of the filament requirements of the unit. This transformer has two 5-volt and two 6.3-volt windings. Each of the 5-volt windings supplies one rectifier tube, while one of the 6.3-volt windings supplies heater power to the entire transmitter with the exception of the push-pull 6A3 stage, which must have a separate winding to allow the use of cathode bias.

Operation

To place the unit into operation it is necessary merely to place the proper crystal in the oscillator stage, throw S_1 to the correct position, depending upon the output frequency desired, switch to the proper plate coil in the HY-69 stage, and tune each stage to resonance as indicated by the meters and the pilot light r.f. indicator. The only trouble which is likely to be experienced is oscillation in the HY-69 stage on 20 meters, the highest frequency at which this stage runs as a straight amplifier. If oscillation occurs, it will probably be traceable to capacity coupling between the antenna

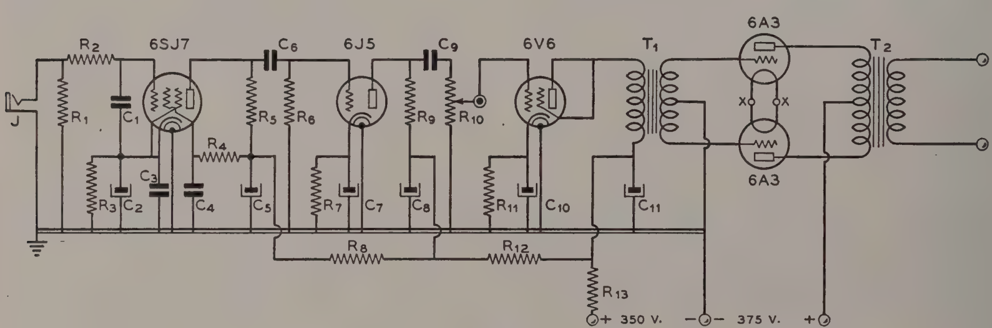


Figure 6.

SPEECH AND MODULATOR CIRCUIT.

C_1 — .0001 - μfd. mica
C_2—10-μfd. 25-volt electrolytic
C_3 — .01-μfd. 600-volt tubular
C_4 — .25-μfd. 600-volt tubular
C_5—8-μfd. 450-volt electrolytic
C_6 — .05-μfd. 600-volt tubular
C_7—10-μfd. 25-volt electrolytic
C_8—8-μfd. 450-volt tubular
C_9 — .05-μfd. 600-volt tubular
C_{10} — 10-μfd. 50-volt electrolytic
C_{11} — 8-μfd. 450-volt electrolytic
R_1—1 megohm, ½ watt
R_2 — 25,000 ohms, ½ watt
R_3—500 ohms, ½ watt
R_4—1 megohm, ½ watt
R_5—100,000 ohms, ½ watt
R_6—500,000 ohms, ½ watt
R_7—1000 ohms, 1 watt
R_8—25,000 ohms, 1 watt
R_9—50,000 ohms, 1 watt
R_{10} — 250,000-ohm potentiometer
R_{11}—600 ohms, 2 watts
R_{12}—15,000 ohms, 20 watts
R_{13} — 2500 ohms, 10 watts
T_1 — Driver transformer for triode - connected 6F6 to class B grids. 3:1 ratio, pri. to ½ sec.
T_2 — 40-watt variable-ratio modulation transformer. Connected to give 3000 - ohm modulator plate-to - plate load with 6000 - ohm r.f. load. (For driver service substitute driver transformer with 3:1 pri. to ½ sec. ratio.)
J — Single - circuit jack

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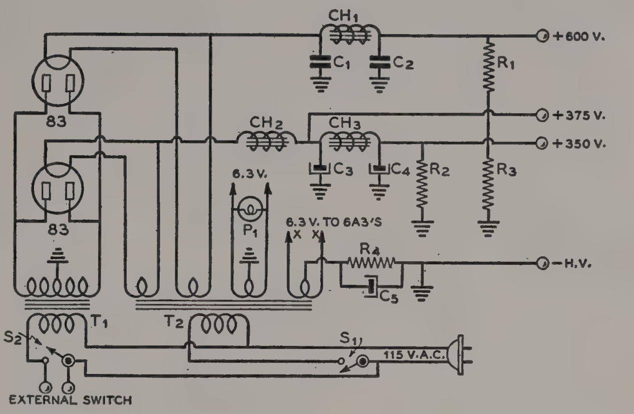
Figure 7.
POWER SUPPLY WIRING DIAGRAM.

T_1—920 v. c.t., 325 ma.
T_2—6.3 v., 5 a.; 6.3 v., 5 a.; 5 v., 6 a.; 5 v., 3 a.
CH_1—30 hy., 100 ma.
CH_2—10 hy., 200 ma.
CH_3—30 hy., 100 ma.
C_1, C_2—4-μfd. 600-volt oil-filled
C_3, C_4—8-μfd. 475-volt elect.
C_5—16-μfd. 150-volt elect.
S_1, S_2—S.p.s.t. toggle
R_1—75,000 ohms, 2 watts
R_2—25,000 ohms, 10 watts
R_3—75,000 ohms, 2 watts
R_4—500 ohms, 10 watts



coupling leads below the chassis and the HY-69 grid circuit, and these leads should be kept well separated. Should oscillation persist, with the antenna leads well separated from the grid circuit wiring and with the transmitter loaded by the antenna, it will be necessary to shield the antenna leads by placing a shield braid over them and grounding the braid to the chassis.

Normal grid current on the HY-69 is 5 milliamperes. The tank circuit may be loaded by the antenna or following stage until the plate current reaches 100 milliamperes.

200-WATT R.F. AMPLIFIER AND MODULATOR

The unit shown in Figure 8 and diagrammed in Figure 9 has been designed specifically to operate in conjunction with the 40-watt r.f. and audio driver unit previously described. Together the two units form a complete 'phone-c.w. transmitter with an output of 200 watts. The r.f. output stage together with its associated modulator and power supply is housed in a 26¼-inch rack cabinet which matches that used for the exciter stages.

The R.F. Amplifier

To provide the best balance between cost and a reasonable amount of power output, the tubes used in the r.f. amplifier are 812's. These tubes are moderate in cost, yet they are capable of producing a 200-watt carrier with a small amount of excitation and a medium-voltage power supply. The input necessary for 200 watts of output is approximately 250 watts (1300 volts at 200 ma.). The photograph of Figure 10 shows clearly the mechanical layout of the stage. The chassis, which measures 17 x 13 x 3 inches, is surmounted by a 14-inch rack-notched panel. The grid coil plugs into a socket near the left edge of the chassis. Between the grid coil and the tubes is located the split stator grid condenser, which

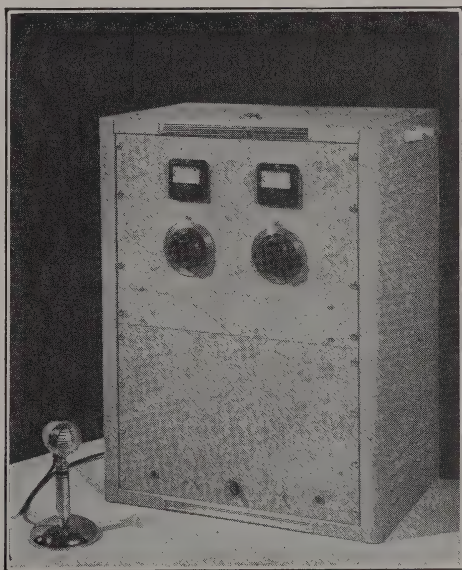


Figure 8.
200-WATT R.F. AMPLIFIER AND MODULATOR.

Inside this cabinet are two chassis, one consisting of a push-pull 812 r.f. amplifier and the other a 1250-volt power supply and a class B 811 modulator. The switches on the lower panel control the filament and plate voltages and disconnect the modulator for c.w. operation. Antenna connections are made to the two stand-off insulators near the top of the right side of the cabinet.

is held $1\frac{3}{8}$ inches above the chassis by spacers to allow its dial to line up with the plate condenser dial. The leads from the grid condenser stators are carried through the chassis to the socket grid terminals by small feedthrough insulators.

To aid in keeping the neutralizing leads short, the neutralizing condensers are placed side by side between the 812's. These condensers are supported from their rear mounting feet by small feedthrough insulators, which also serve to carry the rotor connection to the grid terminals at the sockets. Connecting the neutralizing condensers directly to the grid terminals, rather than to the grid condenser above the chassis, reduces the length of lead which is common to both the neutralizing and tank circuits, thus aiding in securing complete neutralization on all bands. When once set, the neutralizing adjustment need not be changed when changing bands.

Coils. Standard manufactured coil assemblies are used in both the grid and plate circuits of the 200-watt amplifier. The plate coil jack-bar assembly has a swinging pickup loop permanently connected to it. This loop is a flat-wound coil designed specifically to permit a good energy transfer to the antenna regardless of the diameter of the plate coil. The grid coupling loops are an integral part of each grid coil, being mounted in the coil plug in such a way that the coupling may be varied by pushing them in or out of the coil. The coupling should be adjusted so that the grid current measures 50 milliamperes with the amplifier loaded.

The manufactured coils available for use in the amplifier grid circuit require more capacity on the 160-meter band than is provided by the 140-$\mu\mu$fd. per section grid condenser, making it necessary to connect a padder condenser permanently across these coils. The padder consists of a small, ceramic zero-temperature-coefficient 25-$\mu\mu$fd. unit which is permanently connected across the 160-meter coil. It is essential that this condenser be of the type indicated, since the ordinary "postage stamp" type of mica condenser will not stand the circulating tank r.f. current without overheating.

Protective Bias. Relay RY is placed in the grid return circuit to allow protective cathode bias to be applied to the 812's when

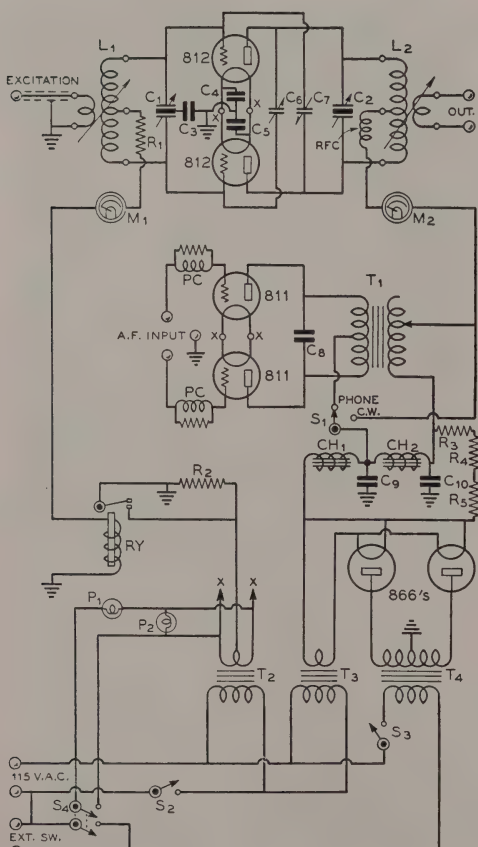


Figure 9.

WIRING DIAGRAM FOR THE 812 AMPLIFIER AND MODULATOR.

C_1—140-$\mu\mu$fd. per section midget variable
C_2—200-$\mu\mu$fd. per section, .100″ spacing
C_3—.002-μfd. mica
C_4, C_5—.004-μfd. mica
C_6, C_7—6-μfd. midget variable, .200″ spacing
C_8—.002-μfd. 5000-volt mica
C_9, C_{10}—4-μfd. 1500-volt oil-filled
R_1—3000 ohms, 20 watts
R_2—500 ohms, 20 watts
R_3, R_4, R_5—100,000 ohms, 1 watt
RFC—5-mhy., 500-ma. choke
L_1—Manufactured variable-link "50-watt" coils. See text for padder on 160-meter coil.
L_2—"500-watt" manufactured coils with variable-link mounting
T_1—125-watt variable-impedance modulation transformer
T_2—6.3 v., 20 a.
T_3—2.5 v., 10 a., 10,000-volt insulation
T_4—2850 v., c.t., 300 ma.
CH_1—8-25 hy. 300-ma. swinging choke
CH_2—15 hy. 200-ma. choke
M_1—0-100 ma.
M_2—0-300 ma.
RY—30-ma. relay
S_1—Single-pole, four-position tap switch (only two positions used). Should have wide spacing between contacts.
S_2—S.p.s.t. toggle
S_3—S.p.s.t. door switch
S_4—S.p.d.t. toggle
P_1, P_2—6.3-volt pilot lamp
PC—Parasitic choke

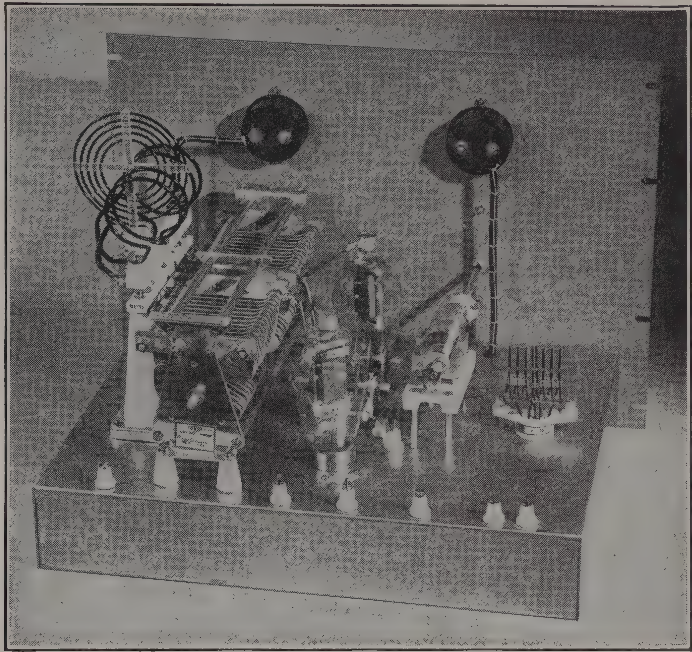


Figure 10.
P.P. 812 R.F. AMPLIFIER.
As with all push-pull amplifiers, symmetry is an important factor in the design of this stage. The plate and tank circuit leads are kept short by sinking the tube sockets below the chassis and mounting the plate coil assembly on tall stand-off insulators.

the excitation is removed. This arrangement allows the exciter to be keyed in the crystal oscillator stage without danger of damaging the final amplifier tubes. It also obviates the necessity for lowering the final amplifier plate voltage when the transmitter is being tuned, since there will always be sufficient bias on the 812's regardless of whether they are receiving grid excitation or not.

The relay is designed to close at a current of 30 milliamperes. When the grid current is less than this amount, the relay contacts are opened and resistor R_2 is cut into the filament center tap circuit, placing sufficient cathode bias on the 812's so that the plate current is held to a safe value.

Modulator and Power Supply

The class B 811 modulator and the 1300-volt power supply for the modulator and r.f. amplifier are mounted on the lower chassis in the rack. Top and bottom views of this section of the transmitter are shown in Figures 12 and 13.

The Modulator. The modulator section of the transmitter needs little comment, since it consists merely of the two 811's and their associated output transformer. The two modulator tubes are located near the left edge of the chassis with the output transformer between them and the panel. The wiring diagram shows parasitic suppressors in the modulator grid leads. These, however, may not be necessary—they are included in the diagram to show where they should be placed in case modulator parasitics should develop. C_8 between the modulator plates reduces high frequency harmonics from the modulator, which cause the signal to "splatter," and this condenser should not be omitted in any case.

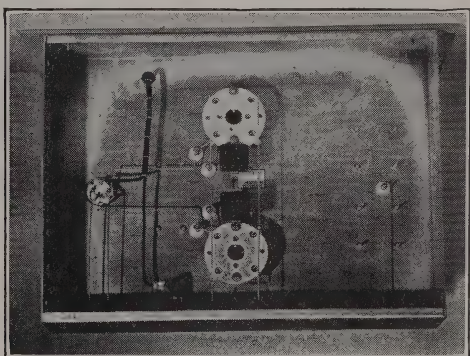


Figure 11.
BOTTOM VIEW OF THE R.F. AMPLIFIER.
The filament leads and most of the grid circuit r.f. wiring are under the chassis. Note that separate feedthrough insulators are used to carry the leads from the socket grid terminals to the grid and neutralizing condensers, thus eliminating common grid and neutralizing leads.

Figure 12.
POWER SUPPLY AND 811 MODULATOR.
The major portion of this chassis is given over to the power supply components. The power transformer is located near the panel to reduce its "leverage" on the panel mounting screws. The modulators are located near the right edge of the chassis with the modulation transformer between the tubes and the panel. Note the safety door switch on the rear drop of the chassis above the right-hand 110-volt connector.

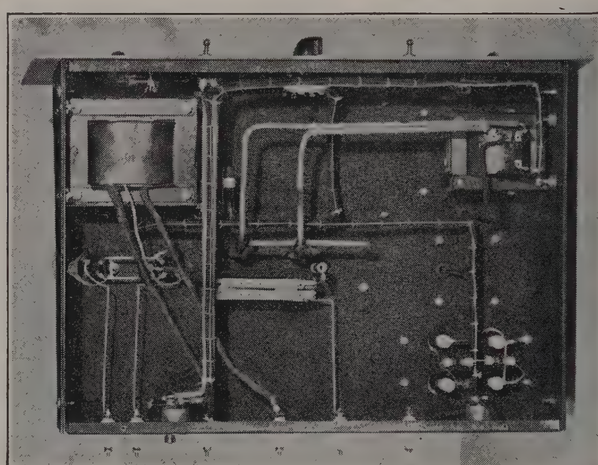


Figure 13.
UNDER THE POWER SUPPLY-MODULATOR CHASSIS.
The 2.5-volt and 6.3-volt filament transformers are located under this chassis. Near the center of the chassis the grid-current-operated safety bias relay may be seen.

The modulator driver transformer is located on the exciter chassis, the correct unit being indicated in the caption under Figure 6. A tapped 125-watt modulation transformer couples the modulators to the r.f. load. The taps on the transformer are adjusted to reflect a 15,000-ohm load on the modulators when working into a 6500-ohm secondary load. Switch S_1, which shorts out the modulation transformer secondary and removes the plate voltage from the modulator for c.w. work, is a ceramic single-pole 4-position tap switch. Only two of the taps on the switch are actually in use—it was chosen because of the wide spacing between contacts.

Power Supply. The power supply section of the final amplifier and modulator unit occupies the center and right-hand portion of the lower chassis. The locations of the various components are plainly visible in Figures 12 and 13.

Of the three switches shown in the power supply wiring diagram, two are on the panel. These are S_2 and S_4. S_2 is placed in series with the primaries of the two filament transformers and controls all of the amplifier fila-

Figure 14.

REAR VIEW OF R.F. AMPLIFIER AND MODULATOR.

Neatly cabled leads between the two chassis aid in giving the unit a finished appearance. The two stand-off insulators near the right edge of the upper chassis are for link connections from the r.f. exciter, while the similar insulators on the lower chassis connect to the audio driver.

ments. S_4 controls the plate voltage to the final amplifier and modulator. S_3 is a safety "door switch" in series with the primary of the plate transformer. This switch is located on the rear drop of the chassis and closes only when the rear door of the rack is closed. It is operated by the long machine screw visible on the inside of the rear door in Figure 14.

The leads marked "external switch" are connected in parallel with the similarly marked leads in the exciter power supply. Closing the plate switch in either the exciter-amplifier section of the transmitter or closing a separate external switch across the leads will turn on the plate power in both sections. Care should be taken to make sure that the side of the external switch line which is connected to the 115-volt supply at the r.f. amplifier-modulator end is connected to the corresponding external switch lead at the exciter end. Since one side of the a.c. supply voltage is connected to the common external switch lead at each unit, care must also be taken in connecting the line voltage to the two units to ascertain that the 115-volt a.c. line will not be shorted. A close inspection of the two diagrams will show the need for observing this precaution.

Three 100,000-ohm, 1-watt resistors, R_3, R_4, and R_5, are used to bleed off the charge in the filter condensers should the power supply be turned off when there is no load being drawn from it. These resistors are included as a safety precaution; they do not serve as a "bleeder" to improve the power supply regulation, since in normal operation no bleeder will be needed because there will always be sufficient load on the power supply, even when the transmitter is being keyed for c.w. operation.

150-WATT C.W. TRANSMITTER

The 150-watt-output c.w. transmitter shown in Figures 15 through 21 has its own self-contained 1400-volt power supply. It may be operated either with its own crystals or excited from an external v.f.o.

The Circuit. The transmitter consists essentially of an 812 amplifier stage excited by

Figure 15.
150-WATT C.W. TRANSMITTER.
This completely self-contained c.w. transmitter delivers 150 watts on 20, 40, or 80 meters.

an 807 tritet crystal oscillator. For use with an external v.f.o. the 807 is made to serve as a buffer or doubler requiring very little excitation.

The 807 tritet stage is conventional in every way. A combination of cathode and grid-leak bias is used, to help keep the crystal current at a minimum. Fortunately, the available power supply voltage is high enough so that the loss of a few volts in the cathode resistor does not reduce the output capabilities of the crystal stage. With the key down the voltage between the 807 cathode and plate is exactly 500 volts.

A double-pole, 3-position tap switch, S_5, takes care of changing the inductance in the cathode circuit when changing to 80- or 40-meter crystals, and also acts to rearrange the circuit slightly for v.f.o. excitation. It will be seen from the diagram that when the switch is in the bottom position, the tuning condenser, C_3, is connected across the whole cathode coil, the connection to the top of the coil being made directly, while the connection to the bottom is completed through ground by means of C_{12} and C_{13}. With the switch in this position an 80-meter crystal may be plugged into XS and the 807 plate circuit tuned to 80, 40, or 20 meters. For 40- or 20-meter operation with a 40-meter crystal, S_5 is thrown to the center position. In this position the top half of L_1 is connected across C_3, while the bottom half is by-passed to ground on each end, thus effectively shorting it for r.f.

To use a v.f.o. to excite the 807, S_5 is thrown to the top position. This by-passes both ends of L_1 (and the 807 cathode) to ground, and also connects the stator of C_3 to the "grid" side of the crystal socket, thereby allowing a coil to be placed in XS in place of the crystal and to be tuned by C_3.

Amplifier Stage. The final-amplifier stage is unusual in respect to most present-day transmitters in its neutralizing circuit. The use of this circuit eliminates some of the troubles so often encountered with single-ended stages at the higher frequencies with the more common split-stator, or "built-out" grid and plate types of neutralizing. Although the neutralizing condenser control is brought out to the panel, the control need not be touched when changing bands, once the coupling between L_2 and L_3 is adjusted to the proper value for each band.

Power Supply. In order to realize the full capabilities of the 812 it must be supplied with 1000 to 1500 plate volts. This amount of voltage is conveniently and economically supplied by a small power transformer and a bridge rectifier using three 5Z3's. With a bridge rectifier, the power transformer center tap may be used to supply a voltage equal to half that obtained from the full supply, and this low voltage is used on the crystal oscillator stage. The main filter choke is placed in the negative lead, where it is common to both the high- and low-voltage sections of the power supply. Additional filter for the low-voltage is provided by an additional choke in the low-voltage positive lead and a pair of 8-μfd. electrolytic condensers in series between the low-voltage positive and ground. The single 2-μfd. 2000-volt condenser across the high-voltage section is adequate filter for the 812 for c.w. work. The high voltage available from the power supply is near 1450 volts under load, the exact amount depending on the line voltage.

Keying. Break-in operation with the transmitter is made possible by keying both stages by the blocked-grid method. The manner in which the keying arrangement works is quite simple, although it is not too evident from the diagram. The blocking bias is obtained by raising the cathode circuits of both tubes up above ground by about 150 volts. It will be seen that the cathode of the 807 and the filament of the 812 are connected together and both leads are run to a tap on the voltage divider, R_{12}. When the key is closed, the section of the voltage divider between the tap and ground is shorted out, thus bringing the cathodes of both tubes back directly to ground in the usual manner. By proper adjustment of the tap on R_{12} the cathode-to-plate voltage on the oscillator may be made to remain constant with the key up or down. This is due to the fact that when the key is up only a portion

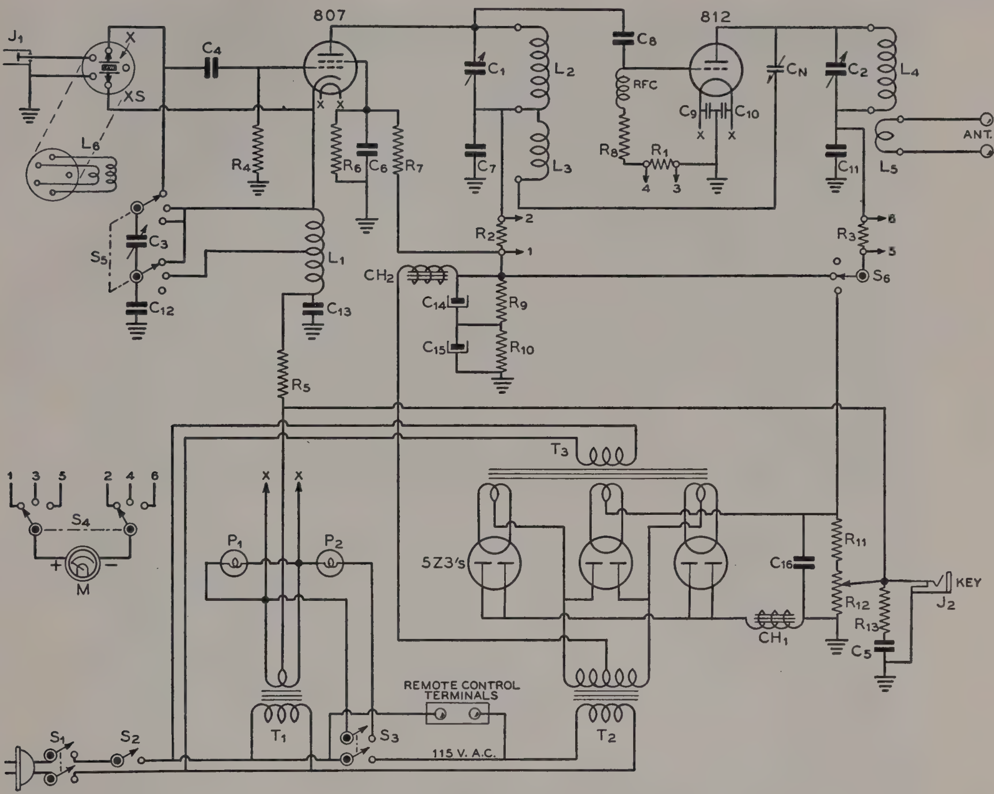


Figure 16.

SCHEMATIC DIAGRAM OF 150-WATT C.W. TRANSMITTER.

C_1—100-$\mu\mu$fd. midget variable
C_2 — 110-$\mu\mu$fd., .078" spacing
C_3—150-$\mu\mu$fd. midget variable
C_4—.003-μfd. mica
C_5—1.0-μfd. 400-volt tubular
C_6—.003-μfd. mica
C_7 — .002-μfd. 1000-volt mica
C_8 — .0001-μfd. 1000-volt mica
C_9, C_{10} — .003 - μfd. mica
C_{11} — .002-μfd. 2500-volt mica
C_{12}, C_{13} — .003 - μfd. mica
C_{14}, C_{15}—8-μfd. 450-volt electrolytic
C_{16}—2-μfd. 2000-volt, oil-filled
C_N — 6-$\mu\mu$fd. midget variable, .200" spacing
R_1, R_2, R_3—100 ohms, 1 watt
R_4 — 50,000 ohms, 1 watt
R_5 — 600 ohms, 10 watts
R_6—25,000 ohms, 10 watts
R_7—15,000 ohms, 10 watts
R_8—10,000 ohms, 10 watts
R_9, R_{10}—50,000 ohms, 2 watts
R_{11}—25,000 ohms, 50 watts
R_{12}—25,000 ohms, 50 watts, with slider
R_{13}—100 ohms, 1 watt
T_1—6.3 v., c.t., 7.5 a.
T_2—1575 v., c.t., 300 ma.
T_3—3 5-volt windings, each 3 amps., high-voltage insulation
S_1—D.p.s.t. "door" switch
S_2—S.p.s.t. toggle
S_3—D.p.s.t. toggle
S_4 — Meter-type tap switch, see text for alterations
S_5 — 3-position, double-pole tap switch
CH_1—Swinging choke 8-40 hy., 250 ma., max.
CH_2—12 hy., 125 ma.
L_1, L_2, L_3, L_4, L_5, L_6—See coil table
RFC—2½ mhy., 125 ma.
M—0-150 ma.
P_1—6.3-v. pilot, green
P_2—6.3-v. pilot, red
J_1 — Automobile-type connector (for link input from v.f.o.)
J_2—Closed-circuit jack
XS—Crystal or grid-coil socket
X — 80- or 40-meter crystal

of the power supply voltage is actually applied to the oscillator tube, and when the key is down the total power supply voltage drops somewhat because of the increased load. By properly adjusting the tap on the voltage divider, it is easily possible to make the key-up and key-down voltage between the 807 cathode and plate have the same value.

The constant-voltage condition on the oscillator may be secured by adjusting the voltage-divider tap so that the voltage between the tap and ground is close to 150 volts. The cor-

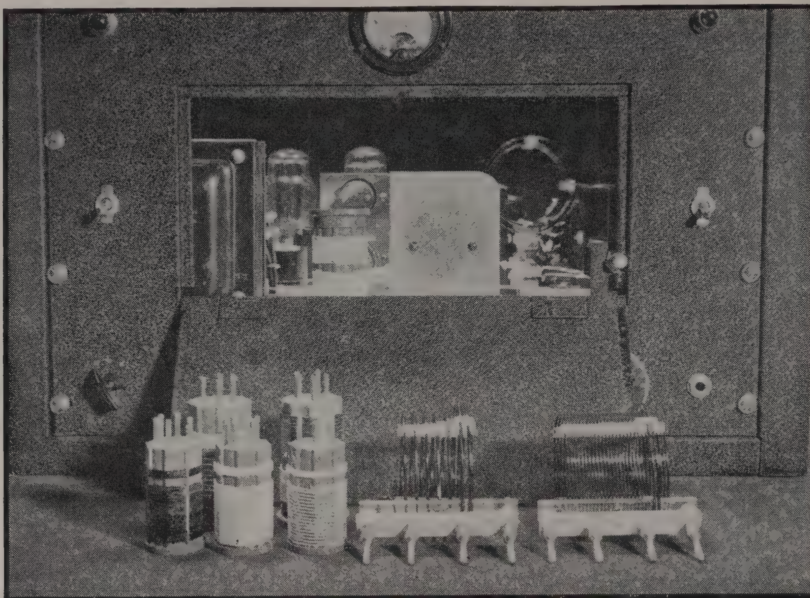


Figure 17.
ILLUSTRATING METHOD OF CHANGING COILS.
Coils are changed through a door in the front panel. A safety switch automatically disconnects the primary a.c. voltage when the door is opened.

rect location of the tap will vary with the loading of the 812 stage, and it is probably best set by actually connecting a voltmeter between the 807 cathode and plate-supply lead and making the adjustment under actual operating conditions. The voltage between the filament and plate of the 812 will also be found to be very constant with this method of keying. When the tap is adjusted for zero change on the oscillator, the 812 plate-filament voltage varies only 50 volts under keying.

Construction. A standard rack-width cabinet is used to house the transmitter. Easy access to the plug-in coils and the crystal is

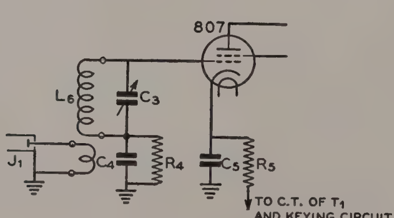


Figure 18.
ALTERNATE CIRCUIT WHICH CAN BE USED WHEN V.F.O. OPERATION IS NOT CONTEMPLATED.

provided by using a panel having a large door in its center. The coils are located so that they may easily be reached through this door. A double-pole "door" switch disconnects both sides of the line from the transmitter when the door is opened, thus eliminating any danger from contact with the a.c. supply or the high voltage. It was deemed advisable to remove the line voltage from the whole transmitter, rather than just from the high-voltage supply, since the 110-volt terminals on the rectifier filament transformer are located where they might possibly be touched when reaching in to change crystals.

To enable the coils to be easily reached through the panel door the rather unorthodox r.f. section construction seen in the photographs is employed. A sheet-metal partition with two 90° bends serves to support both r.f. tubes, and at the same time shields the stages from each other. The partition is 4 inches high. It measures 2½ inches along the side which supports the 807, 5 inches along the long side between stages, and 3 inches along the 812 side. The long side of the partition is located 8 inches from the left edge of the 17 x 12 x 13-inch chassis. Several spade bolts along the bottom edge of the partition serve to hold it firmly to the chassis. The 807 socket is located near the bottom of its side of the

shield partition, to allow space for the crystal socket above the tube. The crystal is thus easily reached through the panel door.

When mounting the 812 socket, it must be remembered that when the tube lies horizontally it must be turned so that the plane of the filament is vertical. Failure to observe this precaution is likely to lead to the untimely demise of the 812 should the filament lean down against the grid. With the filament plane vertical, however, no trouble will be experienced with the horizontal type of tube mounting.

The shaft from S_5, the oscillator cathode-coil bandswitch, extends up through the chassis so that the knob occupies a position alongside the base of the 807. As this switch is used only when changing crystals or when changing from crystal to v.f.o., it is no inconvenience to have the switch behind the panel door, rather than on the panel itself.

The plug-in coils are located so as to be reached easily through the panel door. The oscillator plate coil is located directly in front of the 807, while the amplifier coil is placed alongside the 812. Rounding off the top corner of the shield partition in front of the 812 prevents scratches when the amplifier coil is being changed. However, there is plenty of room

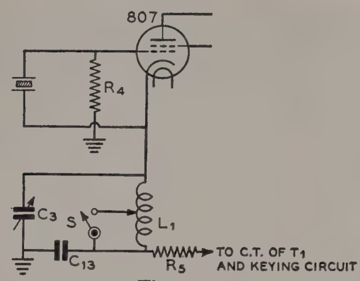


Figure 19.
ALTERNATE CIRCUIT WHICH CAN BE USED WHEN TRANSMITTER IS TO BE USED EXCLUSIVELY WITH AN EXTERNAL V.F.O.

Figure 20.
INTERIOR CONSTRUCTION OF 150-WATT TRANSMITTER.
The location and function of the various parts are covered in the text.

between the shield and the edge of the door to get the coil in and out through the hole in the panel without difficulty.

Most of the power-supply components are mounted above the chassis along the left side and across the rear. The power transformer occupies the left front corner of the chassis; placing it near the front reduces the turning moment on the panel if the transmitter is later to be panel-supported in a large rack. Directly behind the power transformer is the three-winding filament transformer which supplies the 5Z3's. The three rectifiers are placed in a line along the rear of the chassis, followed by the swinging choke, CH_1, and the 6.3-volt filament transformer for the 807 and 812. The high-voltage filter condenser, C_{16}, the bleeder resistors, and the low-voltage filter, C_{14}-C_{15}-CH_2, are in convenient position below the chassis.

A glance at the under-chassis photograph will reveal the location of most of the parts placed in this section of the transmitter. However, due to the angle from which the picture was taken, the shield between the 807- and 812-stage under-chassis components does not show up particularly well. This shield is 9 inches long by 3 inches high and is located directly below the long side of the above-chassis partition. One end of the shield is placed against the front drop of the chassis, the space at the rear being used to allow the power supply wiring to pass back and forth between the ends of the chassis.

A small feed-through insulator is used to carry the lead from L_3 to the neutralizing condenser through the shield. The lead from the neutralizing condenser to the 812 plate runs directly to the bottom of the feed-through insulator which serves to carry the lead from the plate to the tank condenser, C_2. Connecting the neutralizing lead to the insulator, rather than to the tank condenser, keeps the inductance common to the tank and neutralizing circuits to a minimum, thus aiding in securing proper neutralization on all bands.

It is necessary to cut down the length of the meter switch to allow it to fit in front of C_2. As supplied by the manufacturer, the switch has enough spacing between sections so that standard-size 2-watt resistors may be mounted across it, but it may easily be cut down to a length just sufficient to meet the leads from the compact 1-watt resistors seen in the photograph. The cutting down process is quite simple: The back switch wafer is removed, the spacers are pulled off the supporting screws and cut down to a length of ⅞ inch, and the switch is reassembled. The excess screw length may be removed by cutting with a hack saw, or by bending the screws until they break.

COIL SPECIFICATIONS

L_1

The section from the tap to cathode has 7 turns spaced to occupy ½ inch. Section from bottom end to tap has 10 turns close-wound. Form is 1" in diameter. Wound with no. 22 d.c.c. wire.

L_2

80 Meters—19 turns of no. 22 d.c.c. close-wound on 1½" dia. form.

40 Meters—13 turns of no. 22 d.c.c. spaced to occupy ⅞" on 1½" dia. form.

20 Meters—7½ turns of no. 22 d.c.c. spaced to occupy ⅞" on 1½" dia. form.

L_3

80 Meters—42 turns of no. 22 enam., close-wound. Spaced 3/16" from L_2.

40 Meters—21 turns of no. 22 d.c.c., close-wound. Spaced 3/16" from L_2.

20 Meters—9 turns of no. 22 d.c.c., close-wound. Spaced ¼" from L_2.
L_3 and L_2 must be wound in the same direction and L_3 located at the ground end of L_2. The spacing between L_2 and L_3 should be adjusted for proper neutralization as described in the text.

L_6

160 Meters—55 turns of no. 24 enam., close-wound on 1" dia. form. Link—8 turns.

80 Meters—35 turns of no. 22 d.c.c. close-wound on 1" dia. form. Link—5 turns.

40 Meters—19 turns of no. 22 d.c.c. spaced to occupy 1¼" on 1" dia. form. Link—4 turns.

The Coils. Data on winding the 807 plate and cathode coils and the grid coils used for v.f.o. operation is given in the coil table. The plate coils used on the 812 stage are manufactured 150-watt articles. As the manufacturer supplies these coils only with the links at the center, it is necessary to move the links to one end for use with the single-ended tank circuit if the capacity coupling to the antenna is to be kept to a minimum. The links are moved by unsoldering their ends from the plugs and cutting under the celluloid link-spacing blocks with a knife. When the link is loose from the coil it is simply slid to one end

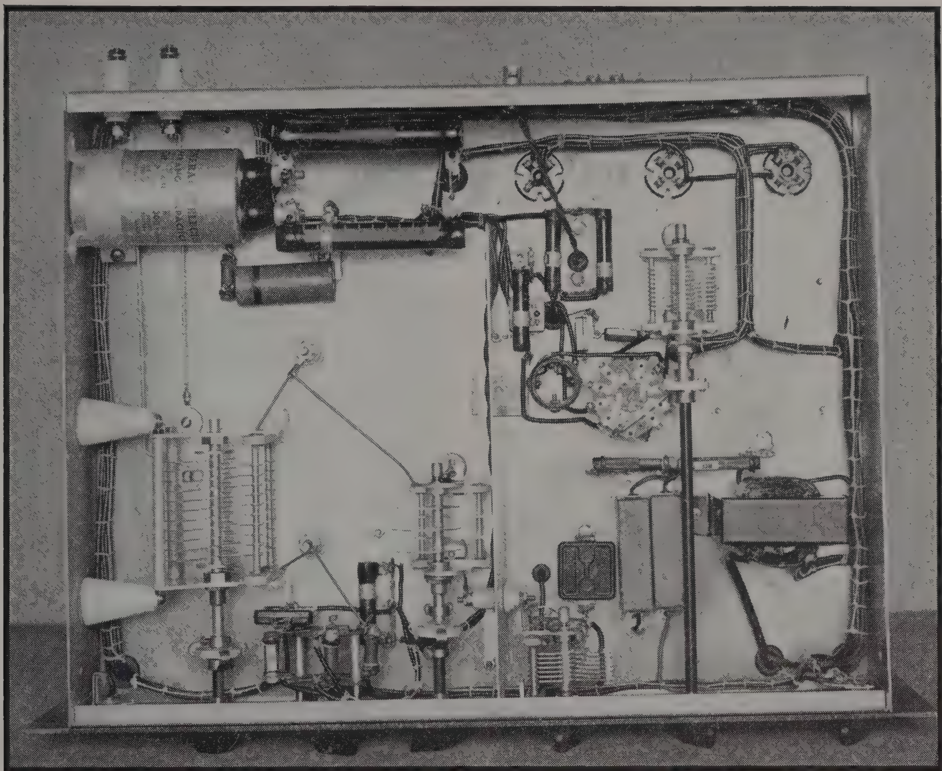


Figure 21.

UNDER-CHASSIS VIEW OF 150-WATT C.W. TRANSMITTER.

At the rear left may be seen the high voltage filter condenser and voltage divider. The variable condenser toward the right rear is used for tuning the 807 cathode coil, which may be seen alongside the wafer type cathode bandswitch. The low voltage filter choke and condensers may be seen near the right front corner of the chassis.

of the coil and cemented in place with Duco cement. The two ends are then reconnected to the plugs. As the center plug is not used in the single-ended circuit, it and the center-tap lead to the coil may be removed, if desired.

To obtain the proper single-ended L/C ratios with the 812 plate coils it is necessary to cut down on the inductance of the manufactured units. Three turns should be removed from the 20-meter coil, 5 turns from the 40-meter coil, and 7 turns from the 80-meter coil.

Tuning Up. The initial tuning of the transmitter is best done on 40 meters, using an 80-meter crystal. The crystal is placed in its socket, and the 40-meter coils are placed in the 807 and 812 plate circuits. Switch S_6 should be set so as to remove the plate voltage from the final amplifier (top position in the diagram), a key is plugged into the keying jack, and S_5 is set to the bottom (80 meter) position. After allowing the filaments to reach operating temperature, S_3 may be closed. If the keying circuit is working properly, there will be no indication of current in any of the three meter switch positions until the key is closed. Closing the key will now give a plate current reading on the 807 stage. This current should be between 60 and 80 ma., depending upon whether the crystal is oscillating or not, and whether the plate circuit is resonated. Placing the cathode tuning condenser, C_3, near maximum capacity should cause the crystal to oscillate, and the meter may be switched to the 812 grid circuit and C_1 tuned for maximum grid current. If all goes well the grid current will be slightly above 30 ma. when C_1 and C_2 are both adjusted for maximum output.

Neutralization may be accomplished by tuning the 812 plate circuit through resonance and observing the drop in grid current. Unless the stage should happen to be neutralized on the first try—which is not likely—there will be a

very pronounced drop in grid current when the plate tank is resonated. Rocking the 812 plate condenser back and forth through resonance with one hand, the neutralizing condenser should be adjusted with the other hand until the variation in grid current is eliminated. If the data in the coil table for L_2 and L_3 has been followed accurately and the stray capacities are about the same as in the original model, neutralization will be obtained when the neutralizing condenser knob is set at 50 on the scale. If more capacity than this is needed for neutralization, L_2 and L_3 should be pushed closer together; if less capacity, they should be separated farther. When the correct spacing between coils has been found, they should be cemented in place with low-loss coil dope.

With neutralization completed, the plate supply may be turned off, S_6 set to the low-voltage (center) position, the meter switch set to read the 812 plate current, and the high voltage again turned on. With the key closed, the 812 plate circuit may then be tuned to resonance, as indicated by the usual minimum plate current point. The minimum plate current should be about 4 milliamperes with the low plate voltage. Opening S_3, switching S_6 to the high-voltage position, and closing S_3, will now put the full power supply voltage on the 812. The minimum plate current should now be approximately 10 milliamperes. The antenna loading should be adjusted so that the plate current under load is approximately 135 milliamperes, which represents an input of nearly 200 watts, and an output of somewhat over 150 watts.

To use a 40-meter crystal "straight through" on 40 meters, S_5 is thrown to the center position and the 807 cathode and plate circuits are again tuned for maximum grid current to the 812. It will be found that when the plate circuit of the 807 is operating on the crystal frequency the plate tuning will be somewhat similar to that obtained with a conventional triode, tetrode, or pentode oscillator. That is, the crystal will pop into oscillation when the plate circuit is tuned to a frequency slightly higher than that of the crystal. The correct setting is the same as with a conventional oscillator—slightly less capacity than the point where the crystal breaks into oscillation.

To operate on 80 meters, an 80-meter crystal is used with the cathode switch set in the 80-meter position. The adjustment of the coupling between L_2 and L_3 should be carried out as described above to secure neutralization at a reading of 50 on the neutralizing condenser scale.

For 20-meter operation either an 80- or 40-meter crystal may be used. The 40-meter crystal is to be preferred, however, since the excitation to the 812 will be rather low with the crystal plate circuit tuned to the fourth harmonic of the 80-meter crystal. The cathode switch must be set at the proper position for the crystal being used, of course. As on the 80- and 40-meter 807 plate coils, the coupling between L_2 and L_3 on the 20-meter coil should be adjusted so that neutralization is obtained at mid-scale on the neutralizing condenser. Once the coupling between these coils is properly adjusted on each band, the neutralizing condenser need not be touched when changing bands. In fact, changing between any two bands can easily be done in less than two minutes, including the time necessary to allow the tubes to warm up after the panel door is closed.

V.F.O. Excitation. To use the transmitter with excitation from a separate v.f.o., the crystal should be replaced with the L_6 coil which matches the output frequency of the v.f.o., and S_5 thrown to the v.f.o. (bottom) position, where L_3 is used to tune L_6. It is preferable to have the v.f.o. output on half the transmitter output frequency, thus doubling in the 807. Although no trouble with oscillation in the 807 stage when running "straight through" on the v.f.o. frequency was experienced in the original transmitter model, perfect shielding between grid and plate circuits is difficult to attain, and doubling is to be recommended. Data for 160-, 80-, or 40-meter grid coils is given in the coil table, so that v.f.o. output on any of these bands may be used.

Antenna Coupling. Since the type of antenna coupling arrangement will depend upon the individual's choice of antenna, no coupling unit is shown. With antennas using an untuned feed line, the feeders may be connected directly to the terminals at the rear of the transmitter, varying the number of turns on the coupling links to secure proper loading. Where an antenna tuner of some type is to be used (see Chapter 20), the link terminals from the coupler may be connected to the transmitter terminals and the coupling adjusted at the antenna end for correct loading.

250-WATT 'PHONE-C.W. TRANSMITTER

The accompanying photographs and diagram illustrate a bandswitching 'phone-c.w. transmitter which is capable of 250 watts input on either 'phone or c.w. on all bands from 80 through 10 meters. The transmitter is complete in every respect in that it includes the entire speech channel and modulator, the antenna tuning network, a click-filtered keying circuit, 'phone-c.w. switch, and, in addition,

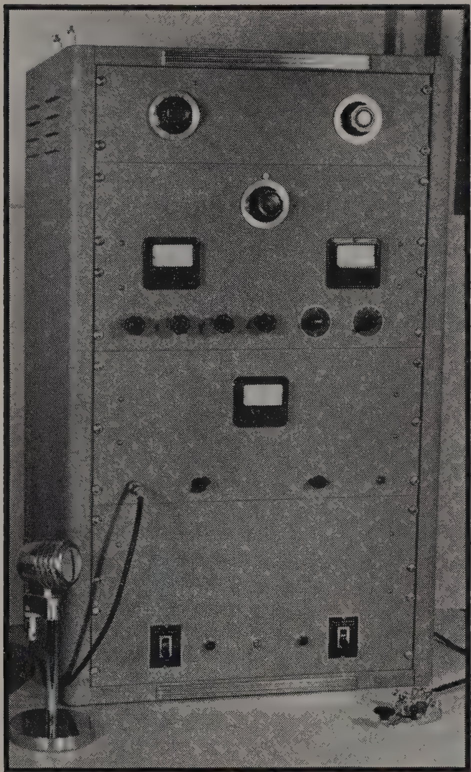


Figure 22.
250-WATT 'PHONE-C.W. TRANSMITTER.
The transmitter is housed in a rack-style cabinet, and it presents a neat and finished appearance. The antenna tuner is at the top, followed toward the bottom by the r.f. section, the speech-modulator, and the power supply.

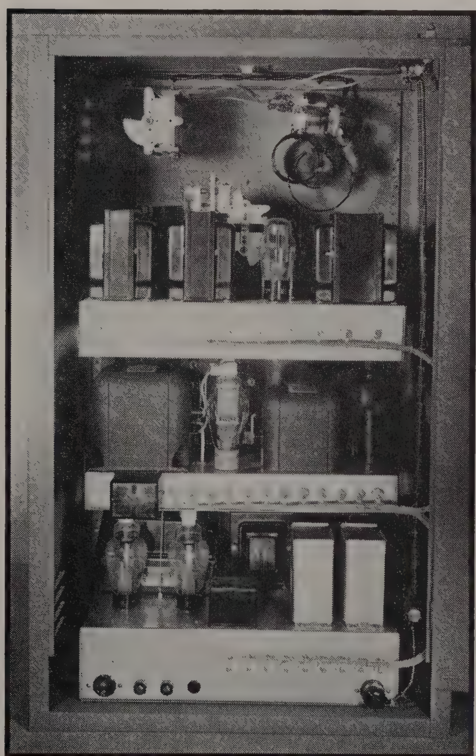
the rig is capable of being controlled and keyed from a distance.

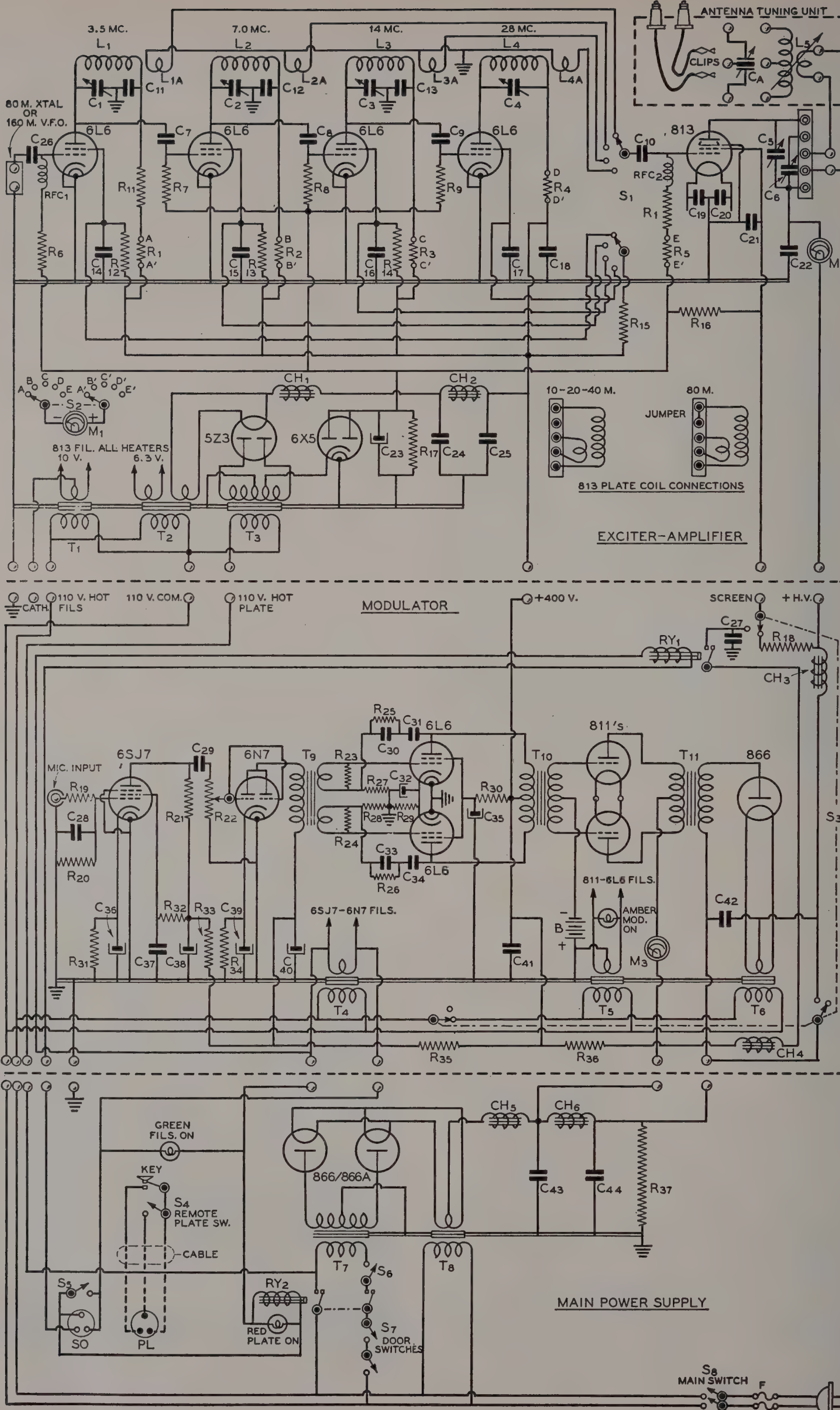
The R.F. Section. Essentially, the r.f. section of the transmitter consists of a 6L6 crystal oscillator stage operating in the 80-meter band followed by three 6L6 doubler stages, and an 813 beam tetrode output stage. If desired, the transmitter may be used with a variable frequency exciter by connecting the v.f.o. output leads across the crystal socket. The v.f.o. output should be at 160 meters, the crystal stage acting as a doubler to 80 meters.

Bandswitching. Excitation to the 813 stage on any band from 80 to 10 meters is obtained by use of non-resonant pickup coils wound around the plate coils of each of the doubler stages. Referring to the circuit diagram (Figure 24) it may be seen that the upper section of S_1 connects the grid condenser of the 813 to the desired doubler. The lower section of S_1 serves to place full screen voltage on the stage being used to excite the final amplifier. Since each of the doubler stages supplies much more output than needed to drive a following 6L6 doubler, there is no need to run these stages at full screen voltage except when they are used to excite the final stage. The last 6L6 (10-meter doubler) is used only to excite the 813 on 10 meters, and for this reason the screen voltage to this stage is removed entirely except when the excitation switch is thrown to the 10-meter position.

It will be noted from the circuit diagram that in the first three exciter stages of the transmitter the tank condenser rotors are

Figure 23.
250-WATT TRANSMITTER—REAR VIEW.
The rear cabinet door is open to show the method of assembly. Interlock switches are provided to disconnect the high voltage when either the top or rear doors are opened. Note how the interconnecting leads between decks are cabled up the side of the chassis.





3.5 MC.
7.0 MC.
14 MC.
28 MC.
ANTENNA TUNING UNIT
CLIPS
80 M. XTAL OR 160 M. V.F.O.
6L6
6L6
6L6
6L6
813
RFC1
RFC2
S1
S2
M1
M2
813 FIL. 10 V.
ALL HEATERS 6.3 V.
5Z3
6X5
CH1
CH2
10-20-40 M.
80 M.
JUMPER
813 PLATE COIL CONNECTIONS
EXCITER-AMPLIFIER
CATH.
110 V. HOT FILS
110 V. COM.
110 V. HOT PLATE
MODULATOR
+400 V.
SCREEN
+H.V.
RY1
CH3
MIC. INPUT
6SJ7
6N7
T9
6L6
6L6
T10
811's
T11
866
S3
6SJ7-6N7 FILS.
811-6L6 FILS.
AMBER MOD. ON
M3
T4
T5
T6
CH4
GREEN FILS. ON
KEY
S4 REMOTE PLATE SW.
CABLE
866/866A
CH5
CH6
RY2
RED PLATE ON
S7 DOOR SWITCHES
S6
SO
PL
T7
T8
MAIN POWER SUPPLY
S8 MAIN SWITCH
F
F

Figure 24.
250-WATT 'PHONE-C.W. TRANSMITTER COMPONENT VALUES.

C_A — 50-μμfd. per section, .171" spacing
C_1, C_2 — 50-μμfd. midget variable
C_3—35-μμfd. midget variable
C_4—15-μμfd. midget variable
C_5, C_6 — 70-μμfd., .070" spacing
C_7, C_8, C_9—.00005-μfd. mica
C_{10}—.005-μfd. mica
C_{11}, C_{12}, C_{13}—.005-μfd. 1000 - volt mica
C_{14} to C_{21} — .003-μfd. mica
C_{22} — .001 - μfd. 5000-volt mica
C_{23} — 25-μfd. 50-volt electrolytic
C_{24}, C_{25} — 4-μfd. 600-volt oil-filled
C_{26} — .0001 - μfd. mica
C_{27} — .25-μfd. 400-volt tubular
C_{28} — .00005-μfd. mica
C_{29} — .02-μfd. 400-volt tubular
C_{30} — .00005-μfd. mica
C_{31} — 0.1-μfd. 400-volt tubular
C_{32} — 50-μfd. 25-volt electrolytic
C_{33} — .00005-μfd. mica
C_{34} — 0.1-μfd. 400-volt tubular
C_{35} — 8-μfd. 450-volt electrolytic
C_{36} — 10-μfd. 25-volt electrolytic
C_{37}—0.25-μfd. 400-volt tubular
C_{38}, C_{40} — Dual 8-μfd. 450-volt electrolytic
C_{39} — 10-μfd. 25-volt electrolytic
C_{41} — 2-μfd. 600-volt oil-filled
C_{42} — .002-μfd. 5000-volt mica
C_{43}, C_{44} — 5-μfd., 2000-volt oil-filled
R_1, R_2, R_3, R_4, R_5—50 ohms, ½ watt
R_6 — 25,000 ohms, 1 watt
R_7, R_8, R_9—100,000 ohms, 2 watts
R_{10}—5000 ohms, 2 watts
R_{11}—2000 ohms, 2 watts
R_{12}, R_{13}, R_{14}—100,000 ohms,2 watts
R_{15} — 15,000 ohms, 10 watts
R_{16}—150,000 ohms, 2 watts
R_{17}—2000 ohms, 10 watts
R_{18}—5000 ohms, 10 watts
R_{19}—50,000 ohms, ½ watt
R_{20}—1 megohm, ½ watt
R_{21}—500,000 ohms, ½ watt
R_{22}—1-megohm potentiometer
R_{23}, R_{24} — 250,000 ohms, ½ watt
R_{25}, R_{26} — 100,000 ohms, ½ watt
R_{27}, R_{28} — 25,000 ohms, ½ watt
R_{29}—200 ohms, ½ watt
R_{30}—7500 ohms, 10 watts
R_{31}—1500 ohms, ½ watt
R_{32} — 2 megohms, ½ watt
R_{33}—50,000 ohms, ½ watt
R_{34} — 1500 ohms, ½ watt
R_{35}—10,000 ohms, 2 watts
R_{36} — 5000 ohms, 10 watts
R_{37}—100,000 ohms, 100 watts
T_1—10 v., 8 a.
T_2—5 v., 3 a.; 6.3 v., 6 a.
T_3 — 1030 v., c.t., bias tap at 30 v.
T_4—6.3 v., c.t., 2 a.
T_5—6.3 v., c.t., 10 a.
T_6—2.5 v., c.t., 10 a.; 7500-v. insulation
T_7 — 3750 v., c.t., 300 ma.
T_8 — 2.5 v., c.t., 7500-volt insulation
T_9—Interstage trans. 1:3 ratio, split secondary
T_{10}—Driver trans. 2.8:1 ratio pri. to ½ sec.
T_{11}—Variable-ratio modulation trans., 125-watt rating
CH_1, CH_2—13 hy., 250 ma.
CH_3—0.8 hy., 300 ma. "splatter choke"
CH_4—15 hy., 85 ma.
CH_5 — 20-5 hy., swinging, 300 ma.
CH_6—12 hy., 300 ma.
RFC_1, RFC_2 — 2½ mhy., 125 ma.
S_1—2-pole, 4-position Isolantite selector switch
S_2—2-pole, 5-position selector switch
S_3—3-pole, 2-position Isolantite selector switch
S_4, S_5—S.p.s.t. toggle
S_6—S.p.s.t. mercury toggle switch
S_7—S.p.s.t. interlock switches
S_8 — D.p.s.t. mercury toggle switch
M_1—0-100 ma.
M_2—0-250 ma.
M_3—0-200 ma.
L_1—30 turns no. 20 d.c.c., close-wound on 1½" dia. form
L_{1A}—9 turns push-back wire over ground end of L_1
L_2—25 turns no. 18 d.c.c. close-wound on 1" dia. form
L_{2A}—9 turns hook-up wire over ground end of L_2
L_3—11 turns no. 20 d.c.c. spaced to occupy 1½" on a 1" form
L_{3A}—5 turns hook-up wire over cold end of L_3
L_4—8 turns no. 12 enam. 1" dia. and spaced to a length of 1½". Self-supporting
L_{4A}—3 turns hook-up wire over ground end of L_4
L_5 — "500-watt" plug-in coils with swinging link
B—4½-volt battery
RY_1—S.p.s.t. 6-volt a.c. coil
RY_2 — D.p.s.t. 6-volt a.c. coil

grounded and the r.f. circuit between the coils and tank condenser completed through mica condensers. On the 10-meter exciter stage, however, the condenser rotor is insulated from ground and condenser and coil by-passed to ground together. This circuit change in the 10-meter stage is made necessary because of the inadvisability of attempting to include a small mica condenser in the tank circuit at such a high frequency.

In the interest of maximum efficiency and compactness, plug-in coils are used in the 813 plate circuit. The four coils are shown in Figure 27. Standard end-linked coils are used on

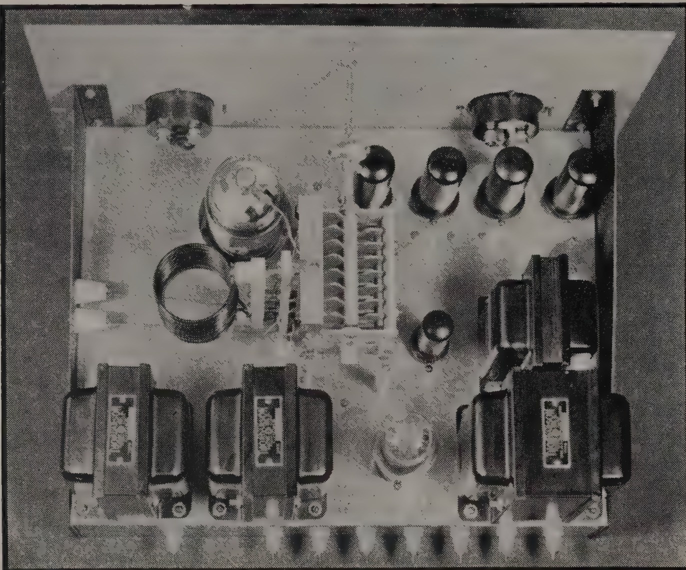


Figure 25.

LOOKING DOWN ON THE 250-WATT R.F. SECTION.

All of the above-chassis components are visible in this view. The four exciter tubes are located in line near the front of the chassis. The 813 and its plate tank circuit occupy the center portion of the chassis, while the power supply components are placed along the rear edge. The bias rectifier tube is alongside the final amplifier plate tank condenser.

the 80- and 40-meter bands, but since the manufactured coils available for use on the 20- and 10-meter bands had too much inductance for use with the high-output-capacity 813, these coils were inexpensively wound to the proper inductance and mounted on the same type jack bar as supplied with the manufactured coils. Data on the winding of the coils for the two high-frequency bands are given under the photograph.

On all bands except 80 meters the tank capacity across the coils is provided by condenser C_5 alone. On 80 meters, however, an extra plug and a jumper on the coil plug bar place the additional 35-$\mu\mu$fd. condenser, C_6, across the coil. The addition of the other 35-$\mu\mu$fd. condenser is necessary to allow a good plate circuit Q to be realized at the lower frequency. Although C_6 is actually a variable condenser, as shown in the diagram, it is permanently set at full capacity and used as a fixed air condenser. The compactness and exact similarity of dimensions of C_6 with C_5 makes it better suited to use in regard to mounting and space requirements than would be a conventional fixed air condenser.

Bias and Power Supply. The exciter power supply utilizes a power transformer which is rated at 515 volts a.c. each side of center tap at 250 milliamperes. This transformer is also provided with a bias tap which delivers 30 volts a.c. for bias purposes. When rectified by a 5Z3 and filtered by a two-section, choke-input filter, the power supply output voltage is 400 volts under load. This voltage is used as plate and screen supply to the exciter stages, as screen supply for the 813 output stage, and as plate voltage to the speech amplifier and driver in the next deck below.

Through the use of a 6X5 as a half-wave bias rectifier, 40 volts of fixed protective bias is made available for all of the transmitter stages. The bias voltage is developed across the load resistor, R_{17}, and is filtered by a single 25-$\mu\mu$fd. electrolytic condenser, C_{23}. Because the current drawn from the bias supply is small, and since the class C operated stages in the transmitter are incapable of operating as grid modulated amplifiers, any small amount of ripple voltage remaining in the bias supply after the small filter is not reproduced in the form of hum modulation on the carrier.

Keying. The transmitter is keyed by means of a built-in keying relay. The leads in series with the 6.3-volt coil of the relay are brought out to the remote-control plug on the back of the transmitter. The keying circuit itself is somewhat unique in that negative bias is applied to the screen of the 813 when the keying relay is open. Inspection of the circuit diagram will show that when the key circuit is open the screen of the 813 is connected to the 40-volt bias supply through the 150,000-ohm resistor R_{16}. But when the keying circuit is closed, the screen is fed from the 400-volt supply through the 5000-ohm resistor R_{36} and the choke CH_4. Condenser C_{27}, which is across the screen circuit when the rig is being operated on c.w., serves to delay the rise and fall of screen voltage as the rig is keyed, and thus give clean keying without clicks or tails.

Since R_{16} has many times the resistance of

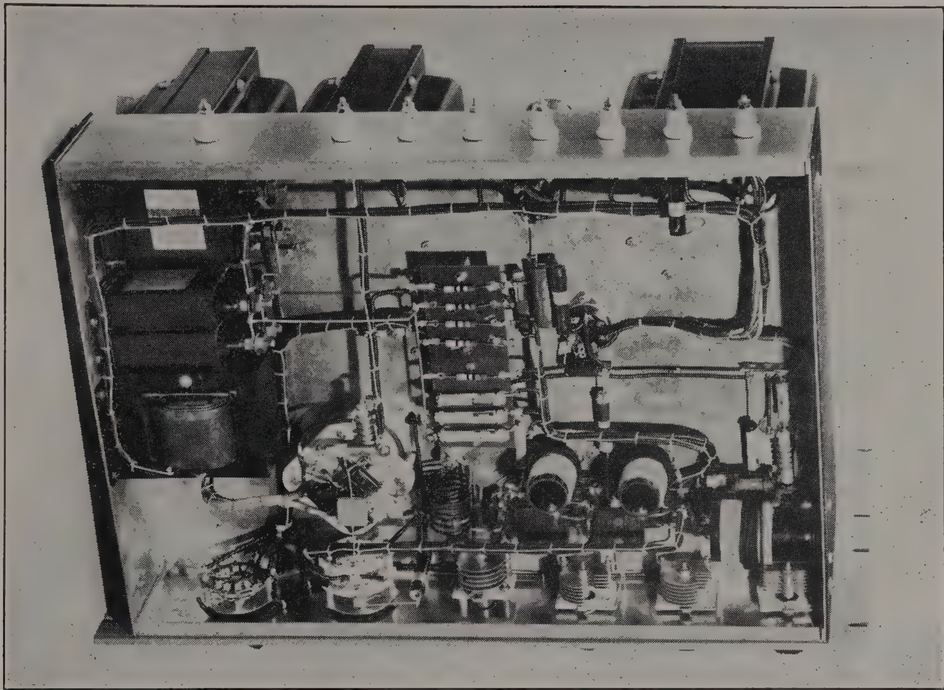
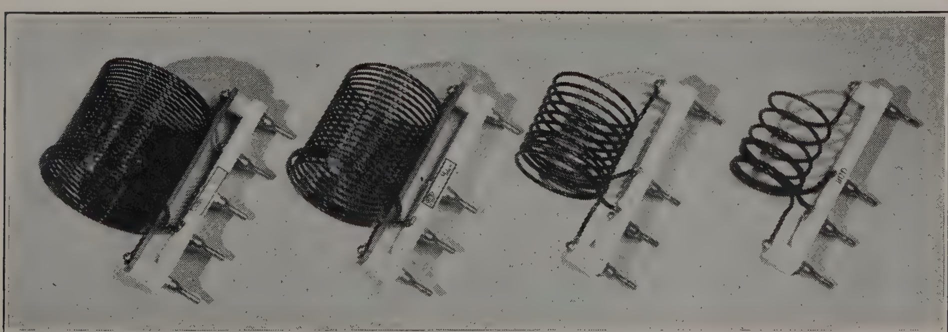


Figure 26.
UNDER-CHASSIS VIEW OF THE 813 TRANSMITTER R.F. CHASSIS.
Most of the wiring is under the chassis. Note that the 813 socket is sunk below the chassis and held in position with the aid of long 6-32 screws and 1-inch hollow spacers. To aid in wiring, the meter resistors and the doubler bias resistors are mounted on a strip at the center of the chassis. Cabling the d.c. leads together aids in giving a neat appearance to the transmitter.

Figure 27.
THE 813 PLATE COILS.
The 80- and 40-meter coils at the left are manufactured units. The 20-meter coil has 9 turns of number 10 enameled wire and is 2½ inches in diameter and 3 inches long. The 10-meter coil is also wound with number 10 enameled wire; it has 5 turns 2 inches in diameter and is 3 inches long. Note the additional plug on the 80-meter coil which serves to connect the extra tank condenser. On each of the higher frequency coils the antenna coupling coil consists of 2 turns of well insulated wire pushed between the turns at the ground end of the plate winding.



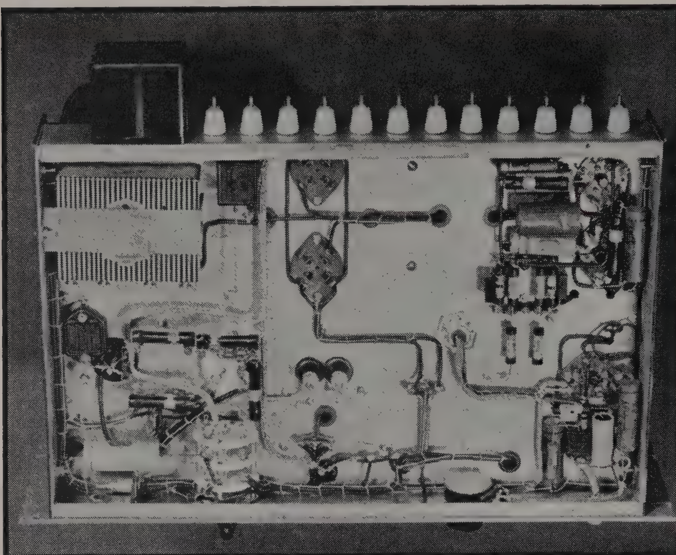


Figure 28.

BOTTOM VIEW OF SPEECH AMPLIFIER AND MODULATOR CHASSIS.

In this photo the speech amplifier section is seen at the right, progressing from bottom (front) to top (rear). The modulator section is at the left. Note the modulator bias battery, the keying relay on the chassis rear drop, and the 'phone-c.w. switch on the front drop.

the bias load resistor R_{17}, no change in bias voltage results when the screen voltage is applied to the 813. The circuit gives exceptionally clean keying at all speeds, since the current through the key is small, and the negative bias when the key is up effectively prevents emission of a "back wave."

The Modulator Channel. The second deck of the transmitter is devoted to the speech amplifier, modulator, and the splatter suppressor circuit. The speech amplifier proper is designed to operate from a conventional diaphragm-type crystal microphone. It starts out with a 6SJ7 high-gain input stage. This is followed by a 6N7, and the driver for the class B stage consists of a pair of 6L6's with degenerative feedback. The speech amplifier has ample gain to operate from any of the common types of high-impedance dynamic and crystal microphones.

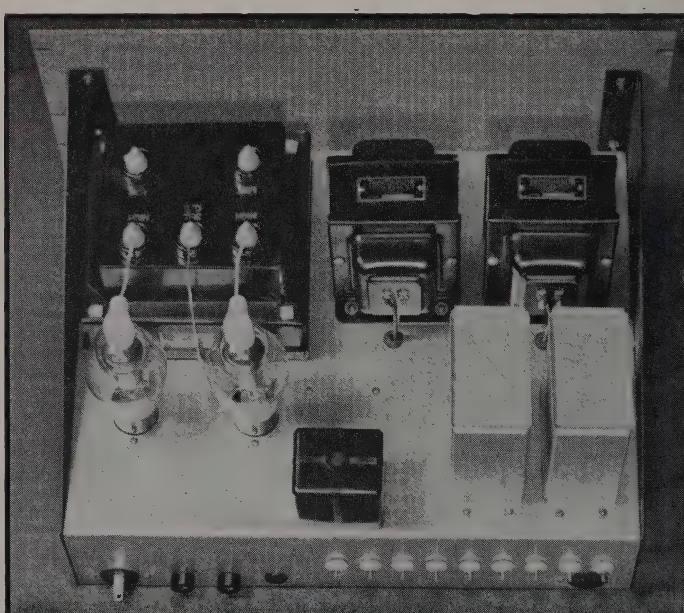


Figure 29.

POWER SUPPLY FOR THE 250-WATT TRANSMITTER.

To keep the height of this deck to a minimum, the power transformer is mounted with the primary side through a large hole in the chassis. The core-clamping bolts are used to hold the transformer to the chassis, and the regular mounting feet are cut off. The two filter chokes, filter condensers, rectifiers, and the keying relay are above the chassis. The rectifier filament transformer and the bleeder resistor are under the chassis.

The modulator stage itself consists of a pair of 811's in class B operating with 1500 volts on their plates and with 4.5 volts of grid bias. Under these conditions of operation, the 811's are easily capable of putting out the 150 to 175 watts of audio power (including the loss in the output transformer and splatter filter and the energy required to modulate the screen of the 813) needed to plate modulate the 813 with 250 watts input. The screen of the 813 is fed modulated plate voltage by means of a drop resistor from the modulated 1500 volts feeding the 813 plate.

The splatter suppressor consists of an 866A/866 rectifier tube in series with the plate lead to the final amplifier, with a 4000-cycle low-pass filter between the rectifier circuit and the modulated amplifier. A complete discussion of the operation of splatter-suppressor circuits has been given in Chapter 8, *Radiotelephony Theory*. Suffice to say here that the rectifier in series with the plate voltage lead eliminates negative-peak clipping, while the low-pass filter attenuates all components of modulation and all components which may be generated by the splatter tube above 4000 cycles.

When the 'phone-c.w. switch on the modulator deck is changed to c.w., the modulation transformer and the splatter suppressor are shorted out, the filament voltage is removed from the 6L6's and the 811's, and the screen circuit of the 813 is removed from the drop resistor going to the final plate supply and connected to the keying relay circuit.

Construction. The transmitter is built into a standard cabinet rack 37 inches high and 14¾ inches deep. The rack has 35 inches of panel space which is apportioned among the various decks as follows: power supply, 10½ inches; modulator 8¾ inches; r.f. section, 10½ inches; antenna tuner, 5¼ inches. The power supply is built upon a 13 x 17 x 4-inch chassis, the modulator upon a 11 x 17 x 2-inch chassis having a bottom plate. The r.f. section and 400-volt power supply are built upon a 13 x 17 x 3-inch chassis, and the antenna tuning network is entirely supported from the panel.

Every effort has been made to keep the 813 plate circuit lead length to a minimum through grouping the tube, coil, and condenser near the center of the chassis. The use of a shield made from a 3-inch coil shield around the base of the 813 above the chassis effectively eliminates any tendency toward oscillation or instability in the final amplifier which might result from capacity coupling between the grid lead within the tube and the plate tank circuit.

Operation. In operating the transmitter it is only necessary to place the proper coil for the desired band in the plate circuit of the output stage, throw the excitation switch to excite the 813 stage from the proper exciter stage, and tune the exciter and final stages to resonance as indicated by minimum plate current. The normal currents on the various stages should be about as follows: oscillator—35 ma.; 40-meter doubler—20 ma.; 20-meter doubler—30 ma.; 10-meter doubler—40 ma.; 813 grid—6–10 ma., depending on band; 813 plate—180 ma., loaded. When the transmitter is tuned up for the first time, the excitation to the 813 on each band should be adjusted to give the required amount of grid current by sliding the coupling coils along the plate coils of each doubler stage.

The antenna tuning network is very flexible in that it is only necessary to change a couple of clips to obtain almost any type of antenna coupling or matching arrangement.

When operating on 'phone, the resting modulator current should be about 45 ma. This current will kick up to about 150 to 175 ma. for normal voice modulation.

300-WATT 'PHONE 500-WATT C.W. TRANSMITTER

The transmitter illustrated in Figures 30–35 is designed for operation on either 'phone or c.w. on all bands from 80 through 10 meters. The final amplifier comprises a pair of 75T's operating at 1500 volts and 200 ma. for 300 watts input on 'phone, and at either 1500 or 1750 volts at 300 to 350 ma. for 450 to 600 watts input on c.w. The transmitter is complete in that it is only necessary to plug in the microphone or key for operation on 'phone or c.w.

R.F. Lineup. The radio frequency portion of the transmitter starts out with a 6L6-G crystal oscillator which may be used either as a straight tetrode oscillator or as a tritet. Crystal switching is provided for the selection of any of three crystals which have been mounted in the transmitter. If desired, the output of a v.f.o. may be plugged into the deck in place of one of the crystals for alternative variable-frequency operation. The selection of straight crystal or tritet oscillator is obtainable by means of a switch on the rear of the exciter deck. Provision is also made on this switch for the selection of a different number of the turns of the cathode coil for tritet operation on different bands.

The buffer amplifier, doubler, or quadrupler is also a 6L6-G which is capacity coupled both to the crystal oscillator stage and to the grid of the following HK-24 r.f. driver stage. The 24 stage operates as a straight neutralized triode amplifier stage on all bands.

The final amplifier is push-pull and uses a pair of 75T's. Split-stator condensers are

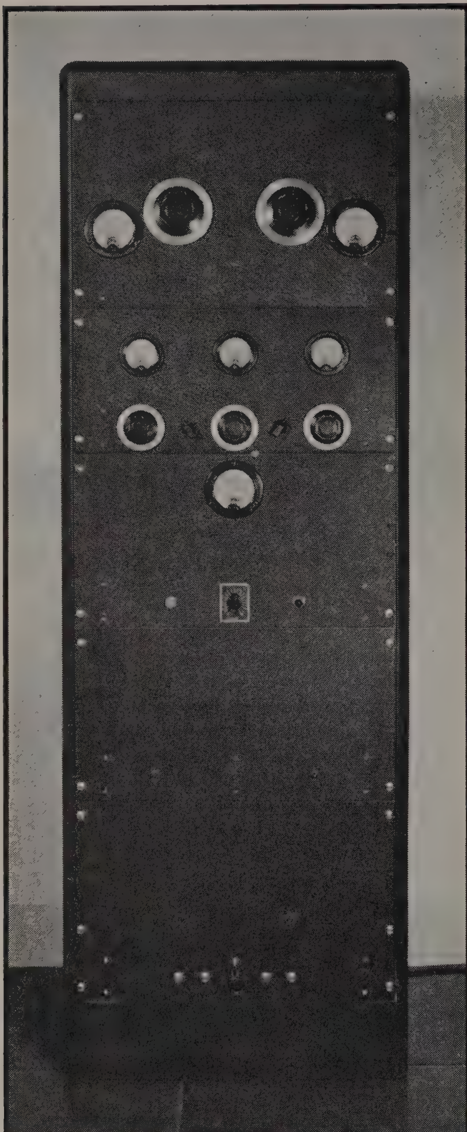


Figure 31.

REAR VIEW OF THE 75T 'PHONE-C.W. TRANSMITTER.

The inter-deck cabling runs just behind the rear upright of the rack cabinet. The a.c. line cord plugs into the right hand socket on the main power supply deck. The 5-wire cable to the operating position also plugs into this deck.

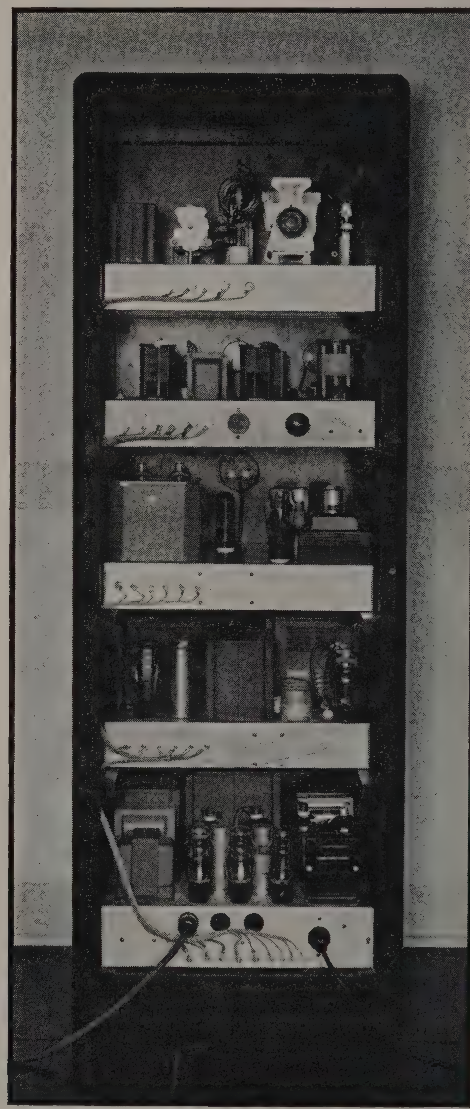


Figure 30.

FRONT VIEW OF THE TRANSMITTER IN ITS ROUNDED-CORNER WOODEN CABINET.

This cabinet contains a complete 300-watt 'phone 500-watt c.w. transmitter for operation on all bands from 80 through 10 meters.

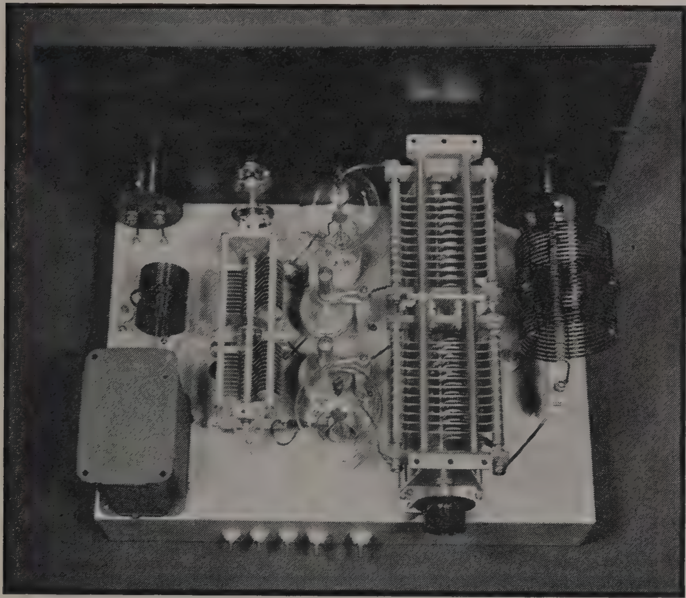


Figure 32.
THE PUSH-PULL 75T FINAL AMPLIFIER DECK.
Note that the filament transformer is mounted directly on the final amplifier deck to insure that the tubes are delivered their full filament voltage. Note also the small parasitic choke in series with the grid of the upper 75T.

used both in the grid and the plate circuits; the rotor of the grid condenser is grounded, and the rotor of the plate tank condensers floats. A commercial type parasitic suppressor of the "6 turns of wire around a 10-ohm resistor" type is inserted in series with the lead from the grid tank and neutralizing condensers to the grid of one of the 75T's. The filament transformer for the tubes is placed on the deck to insure rated filament voltages at the tubes. Grid and plate circuit milliammeters are mounted on the panel and are permanently connected in their respective circuits.

The Audio System. The speech amplifier and modulator of the transmitter is completely self-contained. In the interests of lowered line drain when operating c.w., the audio portion of the transmitter is automatically turned off when the 'phone-c.w. switch is placed in the c.w. position.

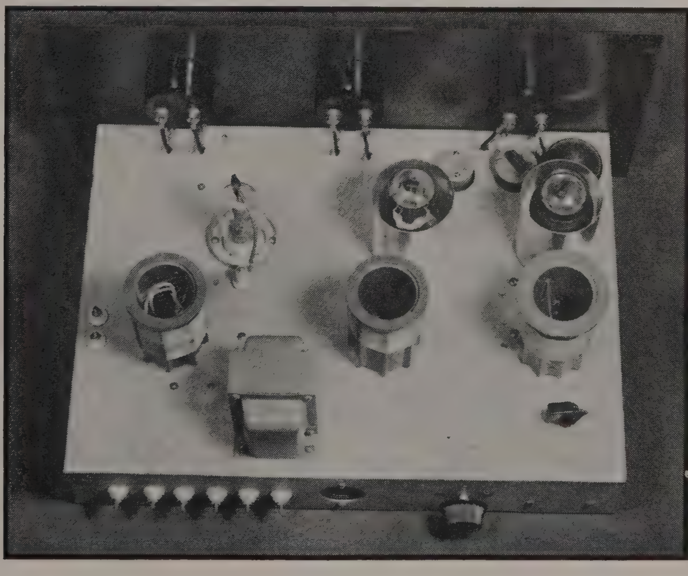


Figure 33.
THE EXCITER DECK.

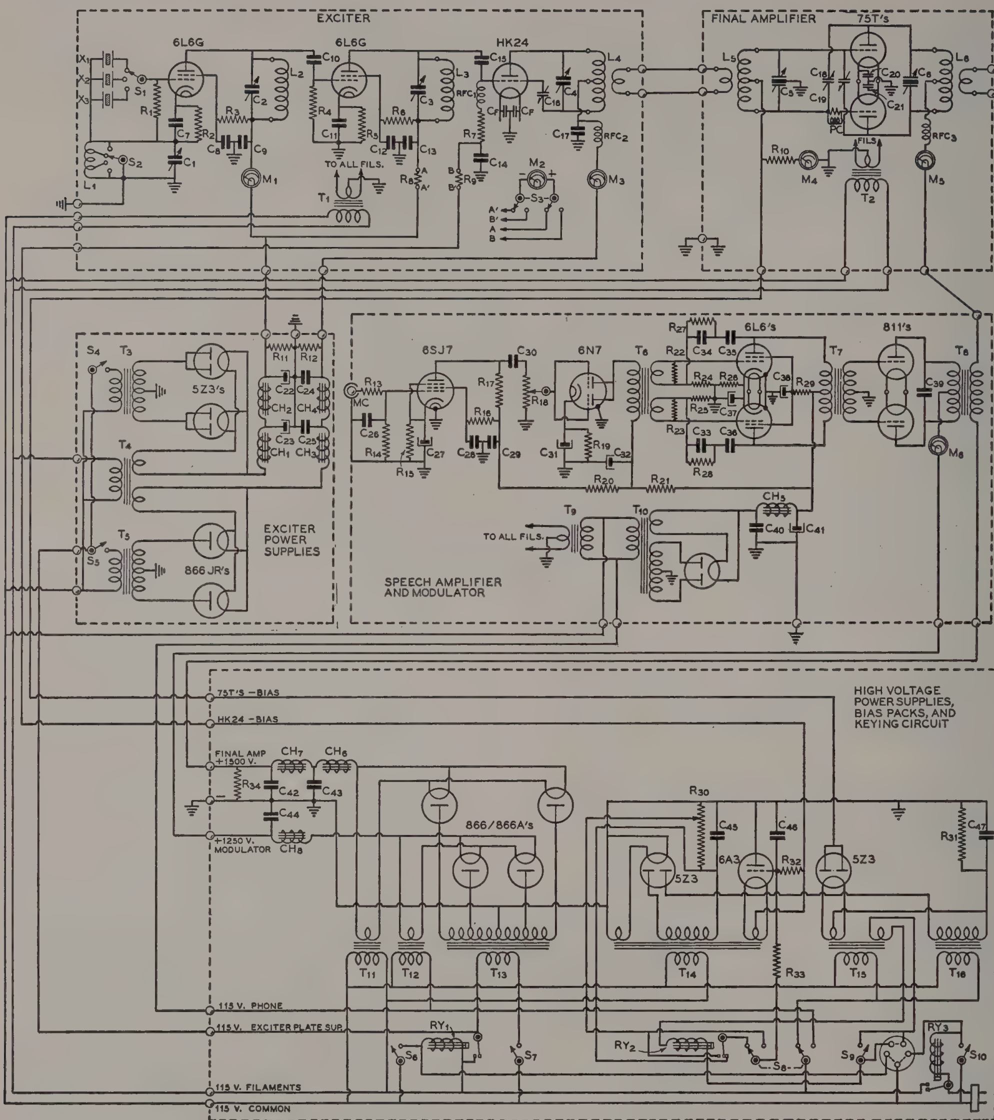


Figure 34.
COMPLETE WIRING SCHEMATIC OF THE 300-WATT 'PHONE 600-WATT C.W. TRANSMITTER.

The actual speech system used is almost identical to that of the 250-watt 813 band-switching transmitter just described, except that the splatter-suppressor system has not been incorporated. The speech lineup consists of a 6SJ7, 6N7, a pair of 6L6's with degenerative feedback, and a pair of 811's in class B. The power supply for the speech amplifier is mounted on the modulator deck and, also, is only in service when the transmitter is being operated on 'phone. It was found necessary to shield the bottom of the chassis in the vicinity of the 6SJ7 input speech stage in order to eliminate all tendency toward r.f. feedback on the 28-Mc. band. As in the 813 transmitter, ample gain is provided for operation from a conventional diaphragm-type crystal microphone.

Power Supplies. Five plate supplies and two grid bias supplies are used in the trans-

C_1—350-$\mu\mu$fd. cathode condenser
C_2, C_3 — 100-$\mu\mu$fd. midget variable
C_4 — 100-$\mu\mu$fd. double-spaced midget variable
C_5—200-$\mu\mu$fd. per section split stator
C_6—200-$\mu\mu$fd. per section .100 spacing
C_7, C_8, C_9—.01-μfd. 600-volt tubular
C_{10} — .0001-μfd. mica
C_{11}, C_{12}, C_{13} — .01-μfd. 600-volt tubular
C_{14} — .002-μfd. mica
C_{15} — .0001-μfd. mica
C_{16} — midget neutralizing condenser
C_{17} — .002-μfd. 1200-volt mica
C_{18}, C_{19} — plate-type neutralizing condensers
C_{20}, C_{21} — .005-μfd. mica
C_{22}, C_{23} — 8-μfd. 450-volt electrolytics
C_{24} — 4-μfd. 1000-volt oil-filled
C_{25} — 2-μfd. 1000-volt oil-filled
C_{26} — .00005-μfd. mica
C_{27} — 10-μfd. 25-volt electrolytic
C_{28}—0.1-μfd. 400-volt tubular
C_{29} — 0.5-μfd. 400-volt tubular
C_{30}—0.02-μfd. 400-volt tubular
C_{31} — 10-μfd. 25-volt electrolytic
C_{32} — 8-μfd. 450-volt electrolytic
C_{33}, C_{34} — .00005-μfd. mica
C_{35}, C_{36} — .01-μfd. 400-volt tubular
C_{37} — 50-μfd. 50-volt electrolytic
C_{38} — 8-μfd. 450-volt electrolytic
C_{39} — .003-μfd. 5000-volt mica
C_{40} — 4-μfd. 600-volt oil-filled
C_{41} — 8-μfd. 450-volt electrolytic
C_{42}, C_{43} — 2-μfd. 2000-v. oil-filled
C_{44} — 4-μfd. 1500-volt oil-filled
C_{45} — 2-μfd. 1000-volt paper
C_{46}—0.15-μfd. 800-volt tubular
C_{47} — 2-μfd. 1000-volt paper
R_1—100,000 ohms, ½ watt
R_2—300 ohms, 10 watts
R_3—50,000 ohms, 2 watts
R_4—75,000 ohms, 2 watts
R_5—300 ohms, 10 watts
R_6—50,000 ohms, 2 watts
R_7—5,000 ohms, 10 watts
R_8, R_9—100 ohms, 1 watt
R_{10}—10,000 ohms, 25 watts
R_{11}—50,000 ohms, 20 watts
R_{12}—75,000 ohms, 20 watts
R_{13}—50,000 ohms, ½ watt
R_{14}—500,000 ohms, ½ watt
R_{15}—1500 ohms, ½ watt
R_{16}—1 megohm, ½ watt
R_{17}—250,000 ohms, ½ watt
R_{18} — 500,000-ohm potentiometer
R_{19}—1000 ohms, 1 watt
R_{20}—50,000 ohms, 1 watt
R_{21}—3000 ohms, 1 watt
R_{22}, R_{23} — 250,000 ohms, ½ watt
R_{24}, R_{25} — 20,000 ohms, ½ watt
R_{26}—200 ohms, 10 watts
R_{27}, R_{28} — 100,000 ohms, 1 watt
R_{29}—5000 ohms, 10 watts
R_{30}—15,000 ohms, 50-watt slider type
R_{31}—50,000 ohms, 20 watts
R_{32}—500,000 ohms, ½ watt
R_{33}—15,000 ohms, 1 watt
R_{34}—100,000 ohms, 200 watts
S_1 — S.p. 3-pos. crystal switch
S_2 — S.p. 4-pos. cathode coil switch
S_3—D.p.d.t. meter switch
S_4 — S.p.s.t. exciter supply switch
S_5—S.p.s.t. 24 supply switch
S_6 — S.p.s.t. main plate on-off switch
S_7 — S.p.s.t. plate-supply safety switch
S_8 — D. p. d. t. 'phone-c.w. switch
S_9 — S.p.s.t. switch in parallel with key
S_{10} — S.p.s.t. main supply on-off switch
L_1 — Tapped tritet cathode coil
L_2, L_3 — Coil suitable for band used
L_4, L_5 — 50-watt type variable link plug-in coils
L_6—500-watt fixed or variable link plug-in coils
X_1, X_2, X_3 — 160-, 80-, or 40-meter crystals
M_1 — 0-50 d.c. milliammeter
M_2—0-100 d.c. milliammeter
M_3—0-150 d.c. milliammeter
M_4—0-100 d.c. milliammeter
M_5—0-500 d.c. milliammeter
M_6—0-500 d.c. milliammeter
RFC_1, RFC_2 — 2.5-mh. 125-ma. chokes
RFC_3 — 2.5-mh. 500-ma. choke
T_1—6.3 v. 5 to 8 a. fil. trans.
T_2—5.25 v. 13 a. fil. trans.
T_3—1050 c.t., 250 ma. plate trans.
T_4 — 35 v. 6 a. windings filament transformer
T_5—2000 c.t. 300 ma. plate trans.
T_6 — Split-secondary push-pull input trans.
T_7—Driver trans. to class B grids
T_8—125-watt multiple-match output transformer
T_9—6.3 v. 12 a. fil. trans.
T_{10}—750 c.t. 120 ma., 5 v. 4 a., (6.3-volt speech fils. if desired)
T_{11}—2.5 v. 10 a. high-voltage insulation
T_{12}—2.5 v. 10 a. high-voltage insulation
T_{13} — 4200/2600/3000 c.t., 300 to 500 ma.
T_{14} — 650 c.t. 40 ma., 5 v. 3 a., 6.3 v. 1 a.
T_{15}-T_{16} — Can be one trans. same as T_{14}
CH_1—5-25 hy. 175-ma. swinging choke
CH_2 — 6-hy. 175-ma. filter choke
CH_3—5-25 hy. 225-ma. swinging choke
CH_4 — 15-hy. 225-ma. filter choke
CH_5 — 6-hy. 175-ma. filter choke
CH_6—5-25-hy. 400-ma. swinging choke
CH_7 — 15-hy. 400-ma. filter choke
CH_8—5-25 hy. 300-ma. swinging choke
RY_1—110-volt a.c. s.p.s.t. relay
RY_2—6.3-volt s.p.-d.t. keying relay
RY_3—110-volt a.c. s.p.s.t. relay
MC — Microphone connector
PC — Parasitic choke

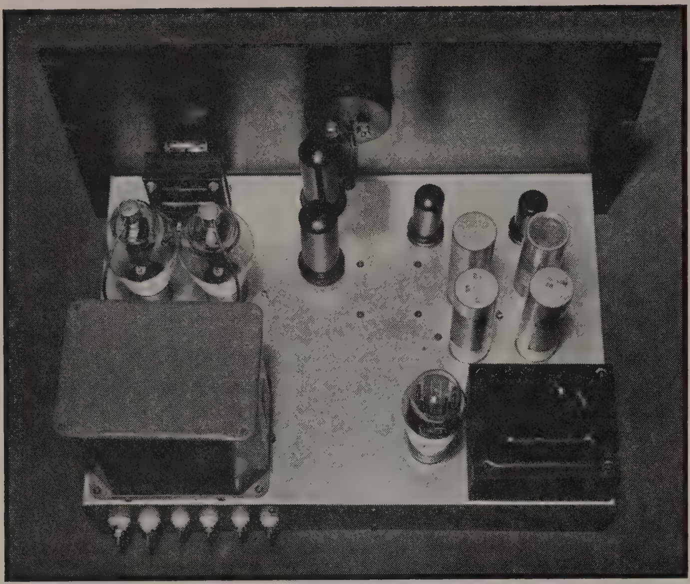


Figure 35.
THE SPEECH AMPLIFIER-MODULATOR DECK.
This entire deck is inoperative, with the power supply off and the filaments unlighted, when the transmitter is operating on c.w.

mitter. These are: 400-volt supply for the speech amplifier, 400-volt supply for the two 6L6's in the exciter, 850-volt supply for the HK-24 buffer amplifier, 1250-volt supply for the class B 811 modulator, the 1500- or 1750-volt supply for the final amplifier, the keyed grid bias supply for the 24, and the minimum bias supply for the final amplifier when operating on c.w.

The speech amplifier supply is operating, as said before, only when the rig is on 'phone. The same is true of the 1250-volt modulator power supply. However, the modulator supply operates from the 1250-volt taps on the transformer that supplies the final amplifier. The filaments of the 866's are turned off when the modulator is not in operation. Inspection of the wiring diagram will show the method of operation. This transformer has taps also which will supply 1750 volts to the final stage instead of the 1500 at which it normally operates.

The bias supply which feeds the grid of the HK-24 is of the voltage regulated type, with a 6A3 regulator tube. For 'phone operation, the grid of the 6A3 connects to a tap on the bleeder of the supply which applies about 100 volts to the HK-24 grid. But when the rig is switched to c.w., the grid of the 6A3 is fed by the keying relay. This relay applies about 500 volts of bias to the grid of the 24 when the key is up (ample voltage to cut off the output of the stage completely), but places the normal 100 volts of bias on the tube when the key is down. A resistor-condenser circuit in series with the grid of the 6A3 serves as a key-click filter. The other bias supply serves merely to bias the tubes of the final amplifier to cutoff when the key is up on c.w. operation. When operating 'phone this supply is not in service (it is cut off by the 'phone-c.w. switch) and the final amplifier operates with straight resistor grid-leak bias.

Remote Control. The transmitter may be completely controlled from a remote position. The 5-prong socket on the back of the power supply deck of the transmitter connects to the three relays in the rig. With a 5-wire cable run to the operating position it is possible to: turn on the filaments, apply the plate voltage, and key the transmitter if it is being operated on c.w. When the rig is being operated on 'phone, the keying circuit is entirely inoperative and it is only necessary to use the filament and plate switches.

Construction. Each of the five decks of the transmitter is built upon a cadmium-plated steel chassis. But the chassis, instead of being supported from the panel as in relay-rack construction, are supported by means of wooden runners on each side of the cabinet. The cabinet itself was constructed by a cabinet maker especially for this transmitter. It is constructed of wood with all corners rounded to give a pleasing appearance. If it is desired to build the transmitter into a manufactured steel cabinet rack of the type used to house the 250-watt 813 transmitter just described, metal shelf brackets could be used on each of the decks to support the chassis from the panels.

400-WATT 10-160 METER PLATE-MODULATED 'PHONE

While the amateur to whom price is no item will naturally want to run a full kilowatt input plate-modulated 'phone when interested in high power, the amateur who is interested in economy will do better to content himself with a 'phone transmitter running in the neighborhood of 600-watts input to the plate-modulated stage. Tubes and modulation transformers for this power are widely available and quite reasonably priced, but when one goes to a full kilowatt the price of these components goes up distressingly. As there is less than 3 db difference (just barely discernible) between a kilowatt and 600 watts input, the cost of the additional power will not be justified in the case of the majority of amateurs.

Hence, for a high-power 'phone transmitter, one delivering about 400 watts of carrier is shown—a very economical size. If one insists upon running a full kilowatt input, it is possible to do so with substantially the same circuit by replacing the 1250-volt power supply with a 1500-volt 400-ma. supply, and the 1900-volt supply with a 2500-volt 400-ma. supply. This will permit the use of an HK254 or 100TH buffer and 250TH's or HK354D's in the modulated amplifier. Slightly greater spacing will be required for the plate tank condenser C_{18}. The 203Z's can be replaced with 822's to deliver sufficient audio power at 1500 volts to modulate fully a kilowatt input on speech waveforms.

Construction of the 400-watt transmitter illustrated obviously is not for the newcomer. And the amateur who has had sufficient construction experience to warrant an attempt at the building of the transmitter will find the illustrations and wiring diagram largely self-explanatory.

The R.F. Exciter. A 6L6G harmonic oscillator driving a 35-T or HK54 neutralized amplifier or doubler forms the exciter portion of the transmitter. The HK54 stage is link-coupled to the grid circuit of the modulated amplifier. The HK54 is first neutralized when working as a straight amplifier on 20 meters. The neutralization will then hold close enough and be sufficiently accurate for operation on all bands. The neutralizing condenser is not disturbed when the stage is used as a doubler.

The Modulated Amplifier. The tubes in the final amplifier "loaf" at between 550- and 600-watts input. While a pair of HK54's or 35-T's could be run at a half-kilowatt input at the plate voltage specified, such input with plate modulation is rather severe and larger tubes will give longer life. HK254's or 100TH's can be run considerably under their rated maximum plate current rating, and very long life can be expected.

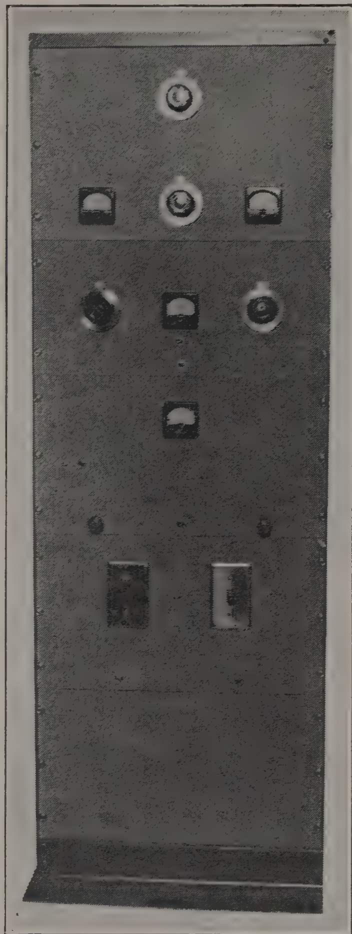


Figure 36.
400-WATT TRANSMITTER.
This 5-foot relay rack contains the complete 400-watt (carrier) radiotelephone transmitter. A pair of class B 203Z's plate modulate a pair of push-pull HK254's.

Sufficient coupling between the buffer and modulated amplifier usually can be obtained with a single turn link around the center of buffer plate and final grid coils. If the grid current to the modulated amplifier runs over 80 ma., the grid tank condenser can be detuned slightly. If it is impossible to obtain 80-ma. grid current on the lower-frequency bands, 2-turn links will be required for those coils.

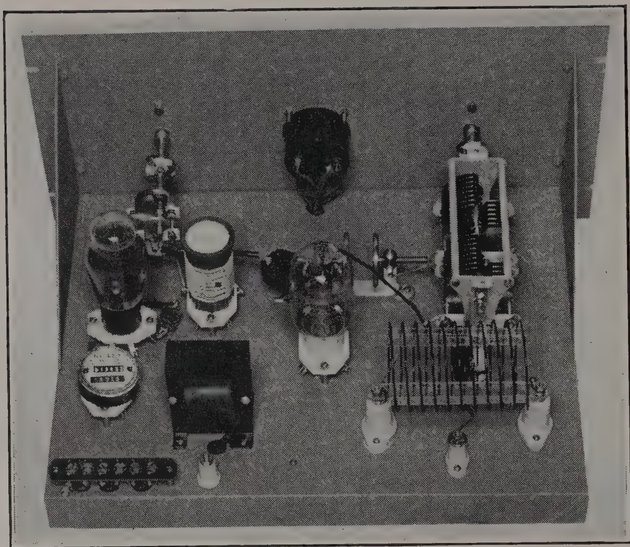


Figure 37.

EXCITER CHASSIS.

The 6L6-G harmonic oscillator and the HK-54 buffer-doubler stages are located on this deck.

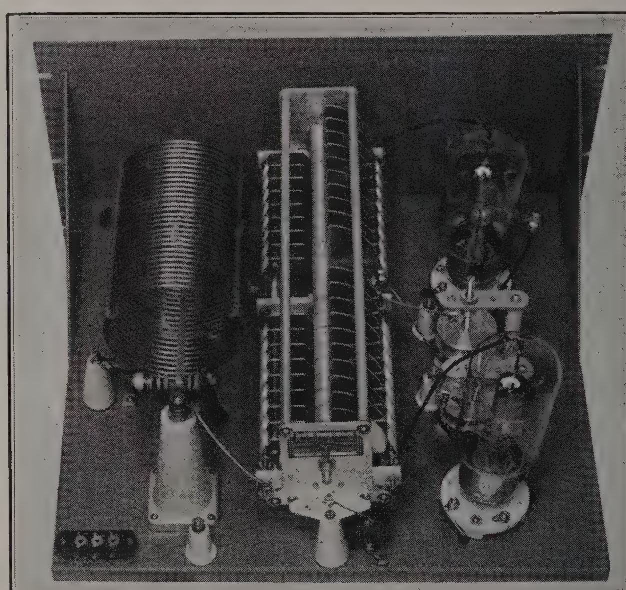


Figure 38.

THE FINAL AMPLIFIER DECK.

A shelf having a narrow lip around it is used to support the final amplifier components. The grid circuit is under the shelf.

To eliminate the need for a more bulky, higher capacity plate tank condenser for 160-meter operation, which would not be advisable for 10-meter operation due to the high minimum capacity, the following expedient is resorted to: the 75-meter amplifier plate coil is made slightly lower Q than optimum. The same coil is then used on 160 meters by shunting a fixed vacuum padding condenser of 50-$\mu\mu$fd. capacity across the tank tuning condenser. This results in a Q slightly higher than optimum for 160-meter operation, but the compromise design of the coil results in operation substantially as satisfactory as would be obtained with separate 75-meter and 160-meter coils.

The Speech System. The speech amplifier-driver and 300-watt modulator are conventional except for the incorporation of automatic peak compression to allow a higher

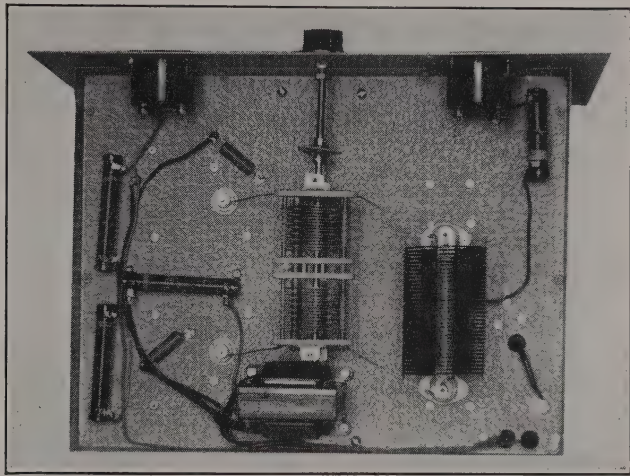


Figure 39.
BOTTOM VIEW OF THE FINAL AMPLIFIER.
The grid coil is shielded from the plate circuit by the metal supporting shelf.

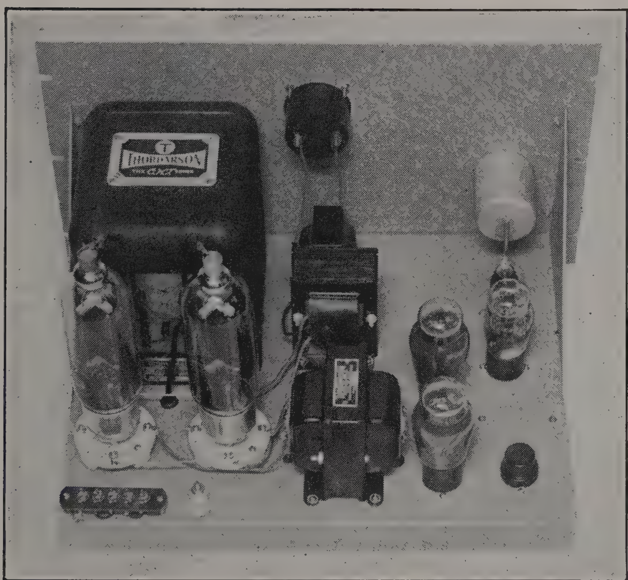


Figure 40.
SPEECH AND MODULATOR DECK.
The entire audio channel is contained in one rack unit. The shield on the back of the panel encloses the input jack, bias cell, grid resistor, etc., and prevents hum pickup.

average percentage of modulation without the danger of overmodulation on occasional loud voice peaks. The delay action (percentage modulation at which compression starts) can be adjusted by means of the potentiometer R_{21}. The modulators are fed from the same 1250-volt supply that furnishes plate voltage to the buffer amplifier.

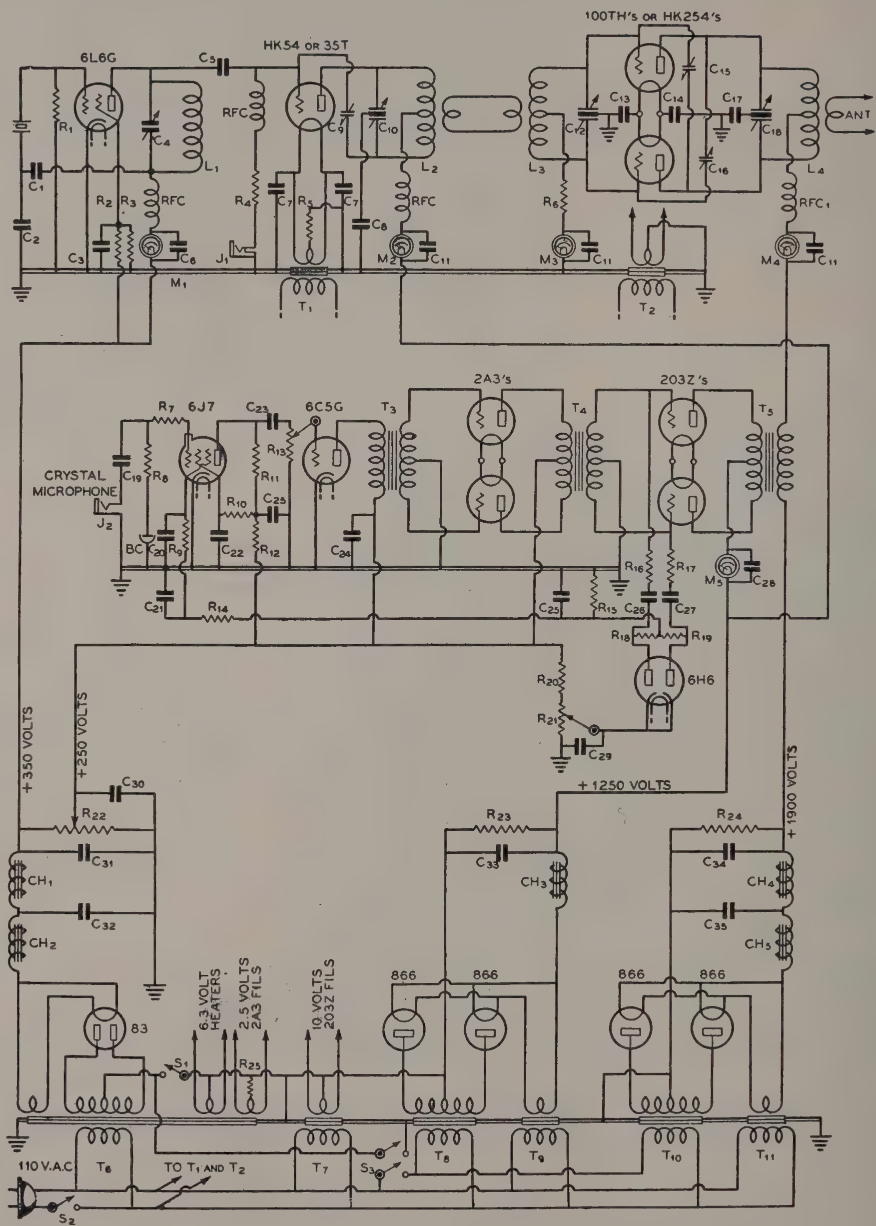
All leads and components in the 6J7 first speech stage should be shielded to prevent grid hum and possible feedback. TZ40's can be substituted for the 203Z's by utilizing 9 volts of fixed battery bias. The tubes will supply sufficient output for complete modulation of 600 watts input when voice is used, though they will not last as long as 203Z's.

The Power Supplies. The 350-volt and the 1250-volt power supplies are built on one chassis; the 1900-volt supply has a chassis of its own. To keep the carrier hum at a very low level, a 2-section filter is used in the 1900-volt supply feeding the modulated amplifier. As the push-pull modulators and the r.f. driver stage are relatively insensitive to a moderate amount of plate supply ripple, a single-section filter suffices for the 1250-volt supply.

While it is desirable to have six meters to facilitate reading of all important grid and

Figure 41.

WIRING DIAGRAM OF THE 400-WATT 'PHONE TRANSMITTER.



CONSTANTS USED IN FIGURE 41

C_1 — .0004-μfd. mica
C_2—.002-μfd. mica
C_3 — .01-μfd. tubular
C_4—50-μμfd. midget
C_5 — .0005-μfd. mica
C_6—.002-μfd. mica
C_7 — .01-μfd. tubular
C_8—.002-μfd. mica, 2500 volts
C_9—Disc type neutralizing condenser
C_{10} — 80-μμfd. per section, 3000-v. spacing
C_{11}—.002-μfd. mica
C_{12} — 80-μμfd. per section, 3000-v. spacing
C_{13}, C_{14} — .01-μfd. tubular
C_{15}, C_{16}—Disc type neutralizing condensers
C_{17} — .0001-μfd. mica, 5000 v.
C_{18}—75 μμfd. per section, 1/4" air gap
C_{19}—.01-μfd. tubular
C_{20}, C_{21}, C_{22}—0.1-μfd. tubular
C_{23} — .01-μfd. tubular
C_{24} — 0.1-μfd. tubular
C_{25} — 0.5-μfd. tubular
C_{26}, C_{27} — 0.1-μfd. tubular
C_{28}—.002-μfd. mica
C_{29} — 1-μfd. paper, 400 volts
C_{30}, C_{31}, C_{32}—8-μfd. electrolytics, 450 volts
C_{33} — 2-μfd., 1500 w. v.
C_{34}, C_{35} — 2-μfd. 2000 w. v.
R_1—100,000 ohms, 1 watt
R_2—10,000 ohms, 10 watts
R_3—50,000 ohms, 2 watts
R_4—15,000 ohms, 10 watts
R_5—300 ohms, 10 watts
R_6—2000 ohms, 50 watts
R_7 — 50,000 ohms, 1/2 watt
R_8—1 meg., 1/2 watt
R_9—250,000 ohms, 1/2 watt
R_{10} — 1 meg., 1/2 watt
R_{11}—250,000 ohms, 1 watt
R_{12}—50,000 ohms, 1/2 watt
R_{13} — 1-meg. tapered pot.
R_{14}—250,000 ohms, 1/2 watt
R_{15}—100,000 ohms, 1 watt
R_{16}, R_{17}—2 meg., 1/2 watt
R_{18}, R_{19}, R_{20}—100,000 ohms, 1 watt
R_{21} — 50,000-ohm pot.
R_{22}—25,000 ohms, 50 watts
R_{23}—75,000 ohms, 100 watts
R_{24}—100,000 ohms, 100 watts
R_{25}—750 ohms, 10 watts
RFC—2.5 mh., 125 ma.
RFC_1 — 2.5 mh., 500 ma.
M_1—0-100 ma. d.c. or meter jack
M_2—0-200 ma. d.c.
M_3—0-100 ma. d.c. or meter jack
M_4—0-500 ma. d.c.
M_5—0-500 ma. d.c.
T_1—5 v. 6 amp.
T_2—5 v. 15 amp.
T_3 — Push-pull input trans.
T_4—Class B input for 203Z
T_5 — 300-watt variable ratio modulation transformer
T_6 — 440 v. each side c.t., 250 ma., and indicated fil. windings
T_7—10 v. 7.5 amp.
T_8 — 1500 v. each side c.t., 300 ma.
T_9—2.5 v. 10 amp., h.v. insulation
T_{10} — 2200 v. each side c.t., 300 ma.
T_{11} — 2.5 v. 10 amp., h.v. insulation
CH_1, CH_2—12 hy., 200 ma.
CH_3—5-20 hy. 300 ma.
CH_4 — 12 hy. 300 ma.
CH_5—5-20 hy., 300 ma.

plate current values simultaneously, it is possible to get by with fewer meters by incorporating metering jacks. Such jacks should be placed in filament return leads rather than in plate leads when the plate potential is over 500 volts. Meters in filament return jacks read combined grid and plate current, and the grid current should be subtracted from the meter reading to determine the actual value of plate current.

Construction. The mechanical construction and lay-out of components can be observed in the various illustrations. All chassis measure 13 x 17 x 1½ inches and have end brackets to strengthen them. All panels are of standard 19-inch width, with heights as fol-

400-WATT 'PHONE TRANSMITTER COIL DATA

BAND	160	80	40	20	10
6L6G PLATE	66 turns no. 22 d.c.c. 1½" diam. close-wound	30 turns no. 20 d.c.c. 1½" diam. 1½" long	15½ turns no. 18 d.c.c. 1½" diam. 1½" long	7½ turns no. 16 enam. 1½" diam. 1¼" long	
BUFFER & FINAL GRIDS	80 turns no. 18 d.c.c. 2⅝" diam. close-wound center tap	36 turns no. 14 enam. 2¾" diam. 8 turns/in. center tap	20 turns no. 14 enam. 2⅝" diam. 5 turns/in. center tap	10 turns no. 14 enam. 2½" diam. 2½ turns per in. center tap	6 turns no. 12 enam. 1¾" diam. 1½ turns per in. center tap
FINAL PLATE	Use 80λ coil shunted by fixed tank condenser (see text)	28 turns no. 10 enam. 4½" diam. 4½ turns per in. center tap	20 turns no. 10 enam. 3½" diam. 3 turns/in. center tap	10 turns no. 10 enam. 3¼" diam. 1½ turns per in. center tap	6 turns no. 10 enam. 2¼" diam. 1 turn/in. center tap

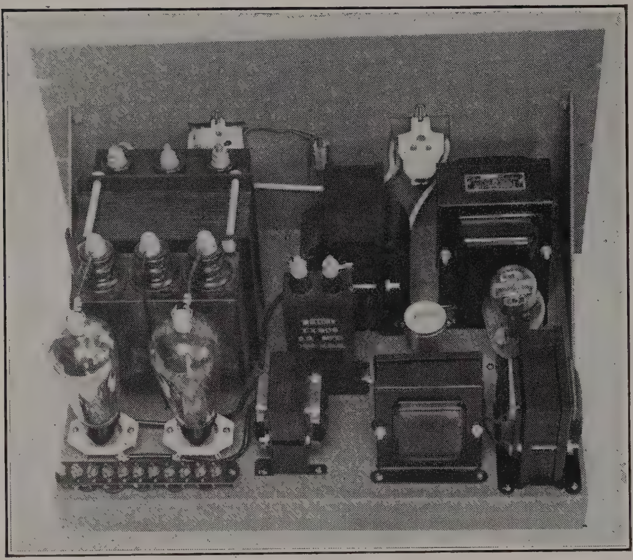


Figure 42.
350- AND 1250-VOLT POWER SUPPLIES.
The low voltage power supply components are located toward the right edge in this rear view.

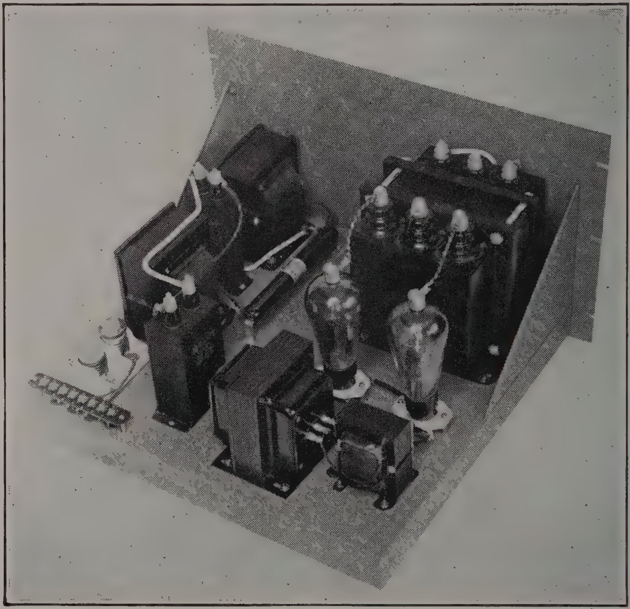


Figure 43.
THE 1900-VOLT POWER SUPPLY.
This power supply feeds the modulated amplifier stage. It has a 2-section filter in order to remove all carrier hum.

lows: final amplifier $12\frac{1}{4}$ inches, exciter $8\frac{3}{4}$ inches, all others $10\frac{1}{2}$ inches.

Operation. Initial tuning of as elaborate and expensive a transmitter as this should preferably be done by an experienced operator who is familiar with tuning and adjustment of high-power 'phone transmitters. General considerations regarding transmitter tuning and adjustments are covered in the transmitter theory chapter. The following meter readings are typical of normal operation:

6L6G cathode current: 35 to 60 ma.

Buffer grid current: 10 to 15 ma.

Buffer plate current: 50 to 75 ma. as buffer; 80 to 100 ma. as doubler.

Final plate current: 300 to 325 ma.

203Z plate current: 75 to 100 ma. resting, swinging up to approximately 200 ma. on voice peaks.

CHAPTER SEVENTEEN

U. H. F. Communication

AN old and still valid definition of *ultra-high frequencies* is: *those frequencies which are not regularly returned to the earth at great distances.* Under this definition, the limit between *high* and *ultra-high* frequencies shifts with the sunspot cycle. From 1935 to 1940, the 30 megacycle band was regularly useful on winter days and on spring and fall afternoons, but through 1945 it is due to become much more erratic. That is not to say, however, that higher frequencies are useless; for the very fact of limitation on distance covered is in itself a blessing for crowded bands, and brings back the old thrill of reaching out to difficult distances. There is good reason for the current trend—or landslide—to these very short wavelengths. In order to promote a better understanding of transmission methods, the several types will be classified and discussed.

Propagation

Direct Communication. *Horizon,* local, or direct point-to-point reception refers to two points between which there is no obstruction to the waves. This might be one mile or two hundred, depending on the altitude of the antennas and the nature of the intervening land.

The distance to the horizon is given by the approximate equation $d = 1.22\ \sqrt{H}$, where the distance d is in miles and the antenna height H is in feet. This must be applied separately to the transmitting and receiving antennas and the results added. However, refraction and diffraction of the signal around the spherical earth cause a smaller reduction in field strength than would occur in the absence of such bending, so that the average radio horizon is somewhat beyond the optical horizon.

There is, however, no sharp discontinuity of the signal at the horizon; that is, an airplane taking off beyond and below the horizon will begin to encounter some signal before reaching an altitude from which the transmitting antenna is actually in sight.

Ground Wave. Because the signal is heard consistently beyond the horizon, the term *ground wave* is usually applied out to 30 or more miles—and much longer when one or both antennas are high. The waves are propagated, presumably, by *diffraction* or dispersion around the curve in the earth's surface in the same way as light is diffracted around a sharp corner. Out to this distance, the transmitting and receiving antennas give best results when both are either vertical or horizontal.

Low Atmosphere Bending. *Pre-skip,* extended ground wave, refracted-diffracted, or low atmosphere bending dx mean essentially the same thing. All refer to distances out to perhaps 200 or 300 miles, in the absence of unusual aurora or magnetic activity. Beams are pointed close to the direct line between the stations. The first two terms refer to the distance but not to the method by which the transmission is accomplished, and presumably differ from the local or ground wave type only because the greater distance is covered as a result of more power, better antennas, or more sensitive receivers.

Low atmosphere bending, on the other hand, in the narrow sense refers to pushing the signal over at the same distance with the aid of a temperature discontinuity or inversion in the lower atmosphere that bends the waves slightly downward, rather than just simple brute force methods implied by the other terms.

When the temperature, pressure, or water-vapor content of the atmosphere does not change smoothly with rising altitude, the discontinuity causes a slight bend in the waves and thus, if the bend is downward, extends the range. Ordinarily this condition is more prevalent at night and in the summer. In certain areas, such as along the west coast of North America, it is believed to be frequent enough to be considered normal. Signal strength decreases with distance and, if the favorable condition in the lower atmosphere covers sufficient area, the range is limited only by the transmitter power, antenna gain, receiver sensitivity, and signal-to-noise ratio. There is no skip

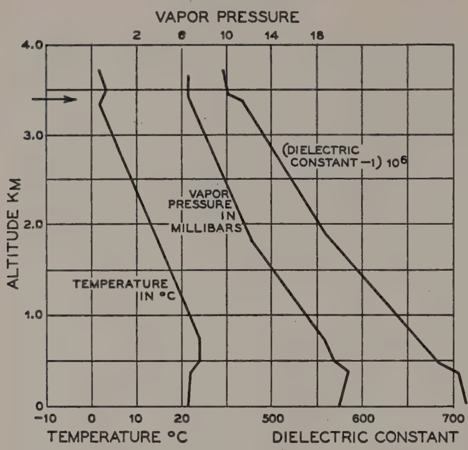


Figure 1.

ILLUSTRATING TYPICAL TEMPERATURE INVERSION AT 3.4 KM.

Air mass boundary heights shown by U.S. Weather Bureau free air data, compared to measured heights from frequency sweep patterns on ultra high frequencies.

distance. Usually, transmission due to this condition is accompanied by slow fading, although fading can be violent at a point where direct waves of about the same strength are also received.

Bending in the troposphere, which refers to the region from the earth's surface up to about 10 kilometers, is more likely to occur on days when there are stratus clouds than on clear cool days with a deep blue sky. The temperature or humidity discontinuities may be broken up by vertical convection currents over land in the daytime but are more likely to continue during the day over water. This condition is in some degree predictable from weather information several days in advance. It does not depend on the sunspot cycle. Like direct communication, best results require similar antenna polarization or orientation at both the transmitting and receiving ends, whereas in transmission via a reflection in the ionosphere (that part of the atmosphere between about 50 and 500 kilometers high) it makes little difference whether antennas are similarly oriented.

Figure 1 illustrates an air mass boundary at 3.4 kilometers, taken from United States Weather Bureau free air data in the vicinity of New York City, at a time when the same height was indicated by ultra-high frequency measurements being made by Bell Laboratories. The arrow points to the inversion or discontinuity in temperature and vapor pressure, and the resulting change in the dielectric constant of the air.

Figure 2 shows typical ultra-high frequency propagation characteristics for a sea water path in the vicinity of New York City, calculated for an air mass boundary at 1500 meters (curve A) and for the earth refracted-diffracted radiation component for ground conductivity 5×10^{-11} E. M. U., and dielectric constant 80 for sea water (curve B) for horizontal and vertical antennas, wave length 4.7 meters (64 megacycles), short doublet antennas, 1 kilowatt power radiated. Most severe fading is generally encountered at such a distance that curves A and B cross, with slow fading at greater distances.

Aurora-Type DX. The same and longer distances can be reached below 60 Mc. during periods of visible displays of the aurora borealis, and during magnetic disturbances. This has been termed *aurora-type dx.* These conditions reach a maximum somewhat after the sunspot cycle peak, possibly because the spots on the sun are nearer to its equator (and more directly in line with the earth) in the latter part of the cycle. Ionospheric storms generally accompany magnetic storms. The normal layers of the ionosphere may be churned or broken up, making radio transmission over long distances difficult or impossible on high frequencies. Unusual conditions in the ionosphere sometimes modulate ultra-high frequency radio waves so that a definite tone or noise modulation is noticed even on transmitters located only a few miles away.

Information is not available as to how high a frequency will be returned by the ionosphere under these conditions, but it is estimated that frequencies from 25 to 100 Mc. may be effected. A peculiarity of this type of propagation of ultra-high frequency signals in the northern hemisphere is that directional antennas usually must be pointed in a northerly direction for best results for transmission or reception, regardless of the direction of the other station being contacted. Distances out to 700 or 800 miles have been covered during magnetic storms, using 30 and 60 Mc. transmitters, with little evidence of any silent zone between the stations communicating with each other. Generally, voice-modulated transmissions are difficult or impossible due to the tone or noise modulation on the signal. Most of the communication of this type has taken place by c.w. or by tone modulated waves with a keyed carrier, and using receivers having an *i.f.* selectivity comparable to that of ordinary commercial high frequency receivers. Because of the association of this type of transmission with magnetic storms, it is assumed that the necessary condition is more likely to occur

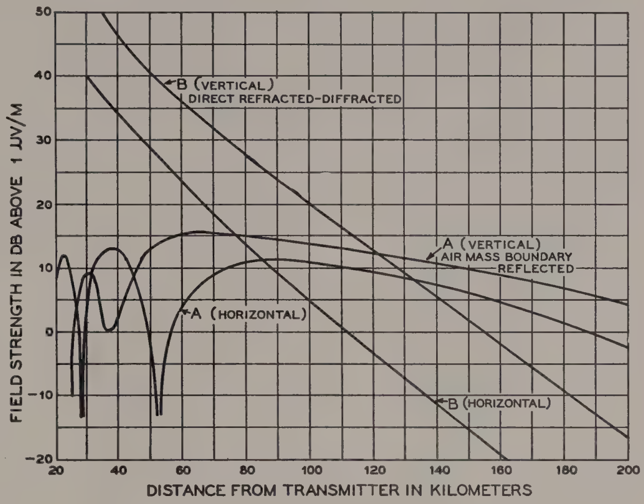


Figure 2.

TYPICAL U.H.F. PROPAGATION CHARACTERISTICS.

Calculated curves for air boundary reflected and earth refracted-diffracted radiation components, in both vertical and horizontal polarization. Short doublet antennas, 1 kw. power radiated, wavelength 4.7 meters, ground conductivity 5×10^{-11} E.M.U., and dielectric constant 80 for sea water. Height of transmitting antenna 42 meters, of receiving antenna 5 meters, air boundary height 1500 meters, effective radius of earth 8500 kilometers.

during or following the sunspot cycle peak, and that this type of transmission may become less frequent during the next 5 years.

Short Skip. The lower of the two more important ionosphere layers is the *E* region. This accounts for 160-meter and broadcast dx at night. Sometimes a *sporadic* condition exists in this layer, the height of which is usually about 110 kilometers (68 miles) above sea level, which will reflect the highest frequency waves that return to the earth. A single hop can be as long as 1,200 miles, or moderately longer at favorable locations or with antennas producing effective low angle radiation (below 3°). Occasionally 1,300 or 1,400 miles can be covered in a single hop, possibly with the help of low atmosphere bending at each end. Sporadic-*E* layer reception may occur at any time, but is much more prevalent from late April to early September in the northern temperate zone, and slightly more likely to occur in the late morning and early evening. The sporadic-*E* layer is spotty, accounting for reception in definite areas completely surrounded by a silent zone, and permitting only a few days of double hop reception during a period of several years. Sporadic-*E* reflections support communication at frequencies up to at least 60 Mc.; reception at as short a distance as 310 statute miles on 56 Mc., in one instance, indicates that the ionization was sufficiently intense so that, theoretically, frequencies as high as 2½ meters (112 Mc.) might have been received erratically at 1,200 miles on that day. At increasing frequencies the silent zone is larger and the reception zone smaller, indicating that the practical limit of sporadic reflections by this layer may be in the vicinity of 80 to 100 Mc.

Since the maximum sunspot activity in 1937–38, the number of hours of reception by reflections from this layer has decreased, at least for frequencies above 50 megacycles and, in general, the skip zone has become larger. The probability of this type of reception may continue to decrease until 1945 or 1946, but improved equipment is making possible 5-meter work, especially, under relatively poor conditions.

When an ionosphere reflection takes place, the polarization of the receiving antenna can be independent of that of the transmitting antenna with equally good results. However, because of the fact that ultra-high frequency antennas are generally placed a number of wave lengths above ground, their vertical plane patterns may contain several angles at which transmission or reception is impossible; a null for a horizontal is at the same angle as a maximum for a vertical antenna located at the same height, and the reverse, thus accounting for widely varying comparisons between the two antenna types, should the waves come in at one of these critical angles.

Beam antennas show some directivity on reception via sporadic-*E* layer reflections, but are not generally as effective as in the low atmosphere bending type of propagation. This has been attributed to the fact that the signal intensity of sporadic-*E* layer transmissions is very high, being comparable to that received within the visible horizon.

Long Skip. The higher of the two major reflecting layers of the ionosphere is the *F* region. This accounts for long-skip signals coming down as far away as 2200 miles in a single hop, with multiple hops common. The silent or skip zone may be around 600 miles

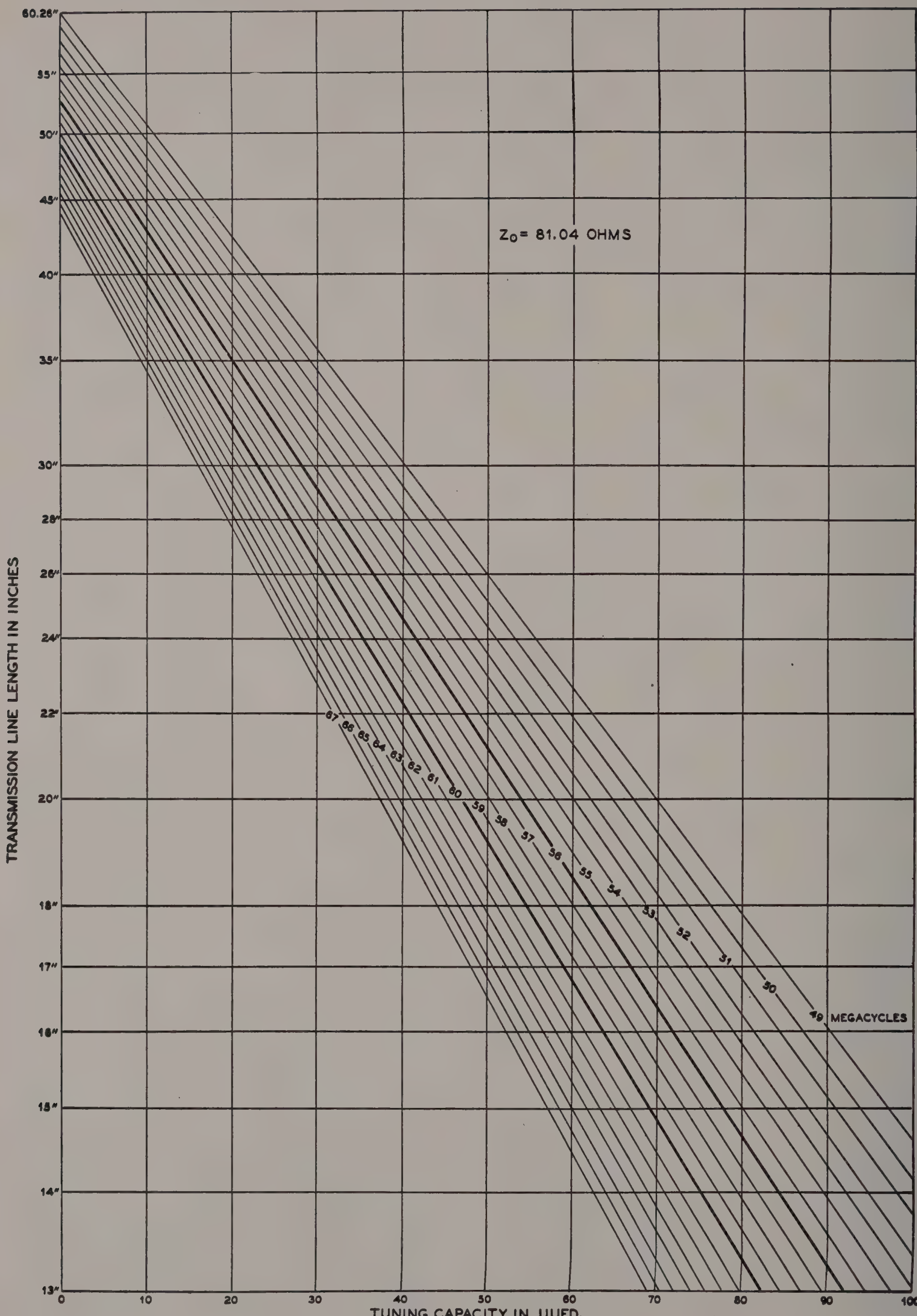

Z0 = 81.04 OHMS
60.26″
55″
50″
45″
40″
35″
30″
28″
26″
24″
22″
20″
18″
17″
16″
15″
14″
13″
TRANSMISSION LINE LENGTH IN INCHES
0
10
20
30
40
50
60
70
80
90
100
TUNING CAPACITY IN µµFD.
67
66
65
64
63
62
61
60
59
58
57
56
55
54
53
52
51
50
49 MEGACYCLES

Figure 3.
CHART SHOWING CAPACITY REQUIRED TO RESONATE SHORTENED LINES OF 81 OHMS SURGE IMPEDANCE.
See text for method of converting to other frequencies and surge impedances. Chart applies directly to coaxial lines and, through conversion (see text), to open-wire lines.

at 30 Mc., and longer at increasing frequencies, when ionization is most intense. On winter days this region accounted for 30 Mc. transmission during the favorable part of the sunspot cycle just passed, except when sporadic-E reflections were present. Observations and ionosphere measurements show that long-distance communication at frequencies as high as 50 Mc. was possible on a few favorable days in 1937 and 1938. Trans-Atlantic 56 Mc. reception for intervals of a few minutes during the favorable part of the sunspot cycle has been reported, but has not been entirely confirmed. Ionosphere and sunspot records suggest that it may be 1947 or so before there is another favorable time for this kind of work.

Trans-Atlantic communication at 30 Mc. was noticeably poorer in early 1941 than in the previous 5 years, and may become almost non-existent for the next 4 years. Transmission southward from the continental United States should be possible at higher frequencies than across the Atlantic.

The silent zone mentioned above has not always been entirely reliable. On days when 30 Mc. transmissions became possible during the past several years, stations have been contacted within the normal skip zone in the morning and afternoon, during which periods it was determined that signals in the northern temperate zone left the transmitter and arrived at the receiver from a southeasterly direction in the morning, and a southwesterly direction in the afternoon, regardless of the location of the stations. This phenomenon has been attributed to reflection from scattered clouds of electrons forming or dispersing in the F region nearer the Equator.

Summary. Ultra-high frequency signals may be heard at distances up to 200 miles or more with fair consistency when antennas are in the clear, and when the receiver has a favorable signal-to-noise ratio. The signal strength within this range depends upon air mass boundary conditions in the troposphere. This type of propagation is effective for frequencies as high as 200–400 Mc.

Scattered reflections during ionospheric or magnetic storms produce an extension of transmission range to at least 500–600 miles, particularly when continuous waves are used at the transmitter and when the receiver passes a narrow band or frequencies. This type of propagation may be limited roughly to frequencies below 100 Mc.

Sporadic-E layer reflections produce extremely loud signals at distances roughly up to 1,200 miles, with a silent zone that is likely to be at least 500 miles at 60 Mc. and 300 miles at 30 Mc., but generally 200 or 300 miles longer. This sporadic condition is unlikely to affect frequencies above 80 to 100 Mc., and is less likely to occur in the winter or during the sunspot minimum which we are now approaching.

For frequencies of 30 to 45 Mc., F layer transmissions in the northern hemisphere are almost entirely confined to the months of August to April over the daylight path. It is unlikely that 60 Mc. signals will be returned by this layer. Conditions for the reception of signals at frequencies between 30 and 50 Mc. are becoming less favorable, and except for sporadic-E layer transmission out to about 1,200 miles, very long-distance reception may become nearly non-existent from 1942 to 1946.

Equipment Considerations

Years ago, tube bases were removed to get down to 100 meters, but experimentation is making $1\frac{1}{4}$ meters (224 Mc. band) as easy as 10 meters was a few years ago. Limits in the use of triode or pentode tubes are being approached, however, which may force further tube and circuit development. Beam tetrode tubes are now available to provide a kilowatt on $2\frac{1}{2}$ meters, and good output on $1\frac{1}{4}$ meters. Triodes are now available to turn out considerable power on $\frac{3}{4}$ meters (400 Mc.). The tuned circuit—the basis of radio—is undergoing changes and may be replaced by *cavity resonance* at microwaves.

Even a perfect circuit must be coupled to something to be useful. A vacuum tube grid presents an apparent low resistance to the tuned circuit at short wavelengths. At 60 Mc., this is about 2300 and 2500 ohms for the 6L7 and 1852, compared with 54,000 for the acorn 954 and 956 and the newer low-priced button tubes, the 9001 and 9003. Normal receiving pentodes such as the type 57 have a relatively low input resistance even at 14 Mc., reducing the effectiveness of the best circuit. With increasing frequency, there is a point for each tube where the output is no larger than the input, adding its shot-effect noise to the signal arriving in its plate circuit. This makes necessary the use of acorn or button type tubes above a certain frequency.

In a properly designed receiver, thermal agitation in the first tuned circuit is amplified by

subsequent tubes and predominates in the output. For good signal-to-set-noise ratio, therefore, one must strive for a high-gain r.f. stage exclusive of regeneration. Hiss can be held down by giving careful attention to this point. A mixer has one-third of the gain of an r.f. tube of the same type; so it is advisable to precede a mixer by an efficient r.f. stage. It is also of some value to have good r.f. selectivity before the first detector in order to reduce noises produced by beating noise at one frequency against noise at another, to produce noise at the intermediate frequency in a superheterodyne or at audio frequencies in a superregenerator.

The frequency limit of a transmitting tube is reached when the shortest possible external connections are used as the tuned circuit, except for abnormal types of oscillation. Generally, amplifiers will operate at higher frequencies than will oscillators. For satisfactory efficiency in an amplifier, it is important to place all tuning condensers so that leads and condenser frame have very little inductance. Otherwise, such leads should be increased to an electrical half wavelength. Wires or parts are often best considered as sections of transmission lines rather than as simple resistances, capacitances, or inductances.

Transmission Line Circuits. At increasingly higher frequencies, it becomes progressively more difficult to obtain a satisfactory amount of selectivity and impedance from an ordinary coil and condenser used as a resonant circuit. On the other hand, quarter wavelength sections of parallel conductors or concentric transmission line are not only better but also become of practical dimensions.

Full quarter wavelength lines resonate regardless of the ratio of diameter to conductor spacing—with due allowance for the length of the shorting disc or bar. Substantial open-end impedance, Z_s, and selectivity, Q, can be built up with lines less than a quarter wavelength, loaded with capacity at the open end, provided that the condenser is an excellent one—preferably copper plates attached to the conductors with no dielectric losses. This is more important, of course, in lines used for frequency control that are lightly loaded. Lines also can be tuned (if not loaded with capacity) by substituting a variable condenser for the shorting bar or disc.

Any unintentional radiation from a coupling link, or resistance coupled into the line, will reduce its effectiveness. Lines that are much shorter than a quarter wave may require considerable capacity to restore resonance; the amount of required capacity can be reduced by using a line with a higher surge impedance —that is, wider spacing for 2-wire lines, or a smaller inner conductor for a given outer conductor of a coaxial line. For greatest selectivity, or oscillator frequency control, the conductor *radius* should be about a quarter of the center-to-center line spacing or, in a coaxial, the inner conductor should be a quarter of the diameter of the outer pipe. For high impedance, ordinarily desired anywhere except for oscillator frequency control, the ratio can be 8-to-1 or higher, thus reducing the necessary loading capacity on short lines.

Very large spacing is undesirable on open wire lines where the shorting bar may radiate so much that the tuned circuit has radiation resistance coupled into it and the impedance is reduced. Preferably, the active surfaces of lines should be copper or silver. A thin chrome plate over copper is also fairly satisfactory, as is an aluminum surface. The conductivity of the center conductor in a coaxial tank is much more important than that of the outer conductor, due to its smaller diameter.

Tuning Short Lines. Tubes hooked on to the open end of a transmission line provide a capacity that makes the resonant length less than a quarter wavelength. The same holds true for a loading condenser. How much the line is shortened depends on its surge impedance. It is given by the equation $\frac{1}{2}\pi fc = Z_0 \tan l$, in which $\pi = 3.1416$, f is the frequency, c the capacity, Z_0 the surge impedance of the line, and $\tan l$ is the tangent of the electrical length in degrees.

The surge or characteristic impedance of such lines can be calculated from the equations: $Z_0 = 276.3 \log_{10} (D/r)$ ohms for 2-wire lines and $Z_0 = 138.15 \log_{10} (b/a)$ ohms for coaxial lines, where Z_0 is the surge impedance, $\log_{10}$ refers to the common logarithm, D and r refer to center-to-center spacing and conductor *radius* of two wire lines, b and a are outer conductor inner diameter and inner conductor outer diameter for coaxial lines. Charts showing characteristic surge impedance for parallel conductors and for coaxial lines may be found in Chapter 20, Figures 12 and 13.

The capacitive reactance of the capacity across the end is $\frac{1}{2}\pi fc$ ohms. For resonance, this must equal the surge impedance of the line times the tangent of its electrical length (in degrees, where 90° equals a quarter wave). It will be seen that twice the capacity will resonate a line if its surge impedance is halved; also that a given capacity has twice the loading effect when the frequency is doubled.

The accompanying chart (Figure 3) can be used to determine the necessary line length or tuning capacity. For 112 Mc., use the 56 Mc. curve but divide the capacity and line length scales by two. That is, if an 81.04-ohm line 30 inches long will tune to 56 Mc. with 28.20-

$\mu\mu$fd. capacity, an 81.04-ohm line 15 inches long will tune to 112 Mc. with a 14.10-$\mu\mu$fd. condenser. Likewise, a 60-inch line of the same impedance will tune to 28 Mc. with 56.40 $\mu\mu$fd. This sounds like a lot of condenser, and can be reduced to 28.20 $\mu\mu$fd. by doubling the line impedance to 162.08 ohms. But, in any event, this circuit will outperform a coil both as to gain and selectivity. The capacities mentioned include circuit capacity; in the case of a mixer preceded by an r.f. stage, this will amount to about 10 $\mu\mu$fd. with acorn tubes, allowing 3 $\mu\mu$fd. for condenser minimum.

Coupling Into Lines. It is possible to couple into a parallel rod line by tapping directly on one or both rods, preferably through blocking condensers if any d.c. is present. More commonly, however, a "hairpin" is inductively coupled at the shorting bar end, either to the bar or to the two rods, or both. This usually results in a balanced load. Should a loop unbalanced to ground be coupled in, any resulting unbalance reflected into the rods can be reduced with a simple Faraday screen, made of a few parallel wires placed between the hairpin loop and the rods. These should be soldered at only one end and grounded.

An unbalanced tap on a coaxial resonant circuit can be made directly on the inner conductor at the point where it is properly matched. For low impedances, such as a concentric line feeder, a small one-half turn loop can be inserted through a hole in the outer conductor of the coaxial circuit, being in effect a half of the hairpin type recommended for coupling balanced feeders to coaxial resonant lines. The size of the loop and closeness to the inner conductor determines the impedance matching and loading. Such loops coupled in near the shorting disc do not alter the tuning appreciably, if not overcoupled. Various coupling circuits are shown in Figure 4.

Frequency Measurement

At ultra-high frequencies, Lecher wires or frames can be used to determine the approximate frequency of an oscillator; a crystal harmonic or receiver oscillator harmonic can then be used for closer measurement. A 10-meter receiver with 1.6 Mc. i.f. will pick up an image 3.2 Mc. from a 10-meter signal. If a 5-meter signal is picked up while the receiver is still tuned to 10 meters, signal and image will be only 1.6 Mc. apart, and the dial setting will be incorrect by one-half of the i.f.

To explain, a 29-Mc. signal would be heard with the receiver oscillator higher in frequency by the amount of the i.f., or 30.6 Mc., with the dial reading 29 Mc. The image would come in when the oscillator is tuned to 27.4 Mc., at which time the dial will read 25.8 Mc. On the second harmonic, however, the dial set at 29 Mc. will place the 30.6-Mc. oscillator harmonic at 61.2 Mc., and bring in signals 1.6 Mc. lower, or on 59.6 Mc. The sub-harmonic of this is 29.8 Mc., or one-half of the i.f. higher than the dial setting of 29.0 Mc.

A 59.6-Mc. signal would also come in as an image when the receiver dial reads 27.4 Mc., or only 1 times the i.f. rather than twice as on the fundamental. At this setting, the oscillator is on 29 Mc., and its second harmonic is on 58 Mc., producing a 1.6-Mc. i.f. by beating against the 59.6-Mc. signal. The above is based on the assumption that the oscillator frequency is higher than the received signal, as is customary in commercial receivers. With a little care, this method can be used to spot bands as well as to place a transmitter in a band with fair accuracy.

Lecher Wire Systems. A Lecher wire measuring system consists of a pair of parallel wires one or more wavelengths long, short circuited at one end to provide a pick-up loop

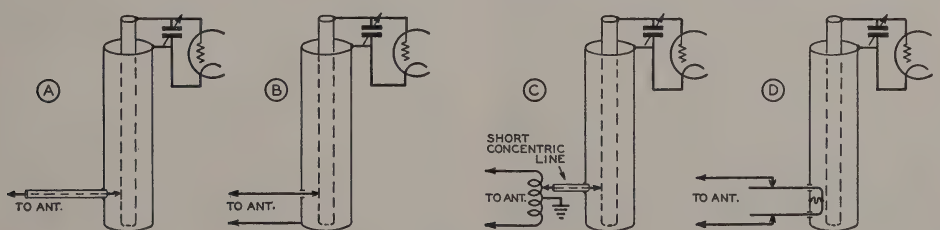


Figure 4.

METHODS OF COUPLING ANTENNA TO COAXIAL RESONANT CIRCUIT.

(A) Coupling a concentric line feeder to a concentric line resonant circuit. (B) Unbalanced method of coupling 2-wire line into a concentric line circuit. (C) Balanced-to-unbalanced method of coupling a 2-wire line to a concentric line resonant circuit. (D) Balanced loop method of obtaining good coupling from 2-wire line to a concentric line circuit.

Frequency (Mc.)	¼ Wave (inches)	½ Wave (inches)
56	52.7	105.5
57	51.8	103.6
58	50.9	101.8
59	50.0	100.1
60	49.2	98.4
112	26.4	52.7
113	26.1	52.3
114	25.9	51.8
115	25.7	51.3
116	25.4	50.9
224	13.2	26.4
226	13.1	26.1
228	12.9	25.9
230	12.8	25.7
400	7.4	14.8
410	7.2	14.4

FREQUENCY VS. WAVELENGTH

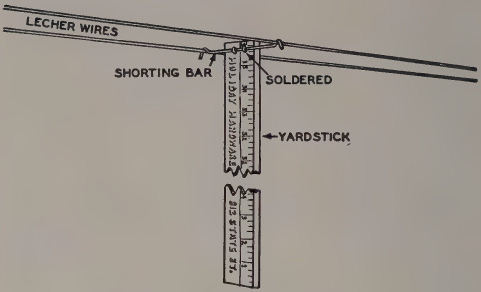


Figure 5.
LECHER WIRE MEASURING EQUIPMENT.

The wires are spaced about 1½ inches and pulled taut. "Bumps" will appear exactly ½ wavelength apart on the wires as the jumper is slid along. The wires may be coupled to the oscillator under measurement by means of twisted line.

which can be coupled to the tuned circuit of a transmitter or receiver. The wires can be no. 12, approximately 1 inch apart. The shorter wavelength units can be stretched on a long wooden framework if no supports or insulators are used in the measuring range.

Energy induced in the parallel wires establishes standing waves of voltage and current along the wire when resonance is established with a shorting bar. The sliding bar (see Figure 5) is moved along the wires until two successive points are located which cause the oscillator under test to draw more plate current or go out of oscillation. The distance between these two points is a half wavelength. This can be converted into meters by multiplying the length in feet by 0.61 (actually 0.6096), or the length in inches by 0.0508. For microwaves, the length in inches is usually converted to wavelength in centimeters by multiplying by 5.08. These factors convert to the metric system and take care of the fact that the points are one-half rather than one wavelength apart. An accuracy of only 1 per cent or so can be expected; receiver or oscillator harmonics should supplement these measurements for greater accuracy.

Lecher Frames. For a quick check of wavelength, any two parallel wires or rods can be used as a quarter wave Lecher *frame*. The open ends can be held near the oscillator while a screw driver or other shorting bar is run down the rods. The oscillator frequency will change and the output will dip when the Lecher frame crosses resonance. This point will give a close approximation of the frequency if half the shorting bar length plus one conductor from the shorting bar to the end near the oscillator is taken as 0.95 of a quarter wavelength. Accuracy to better than 3 per cent can be expected with this system.

Receiver Theory

So long as small triodes and pentodes will operate normally, they are generally preferred as u.h.f. tubes over other receiving methods that have been devised. However, the input capacity of these tubes limits the frequency to which they can be tuned. The input resistance, which drops to a low value at very short wavelengths, limits the stage gain and broadens the tuning. The effect of these factors can be reduced by tapping the grid down on the input circuit, if a reasonably good tuned circuit is used.

A mixer or detector can have a gain only of about one-third of that for the same tube used as an r.f. amplifier, so that for gain and principally for satisfactory signal-to-set-noise ratio, a good r.f. stage is advisable. The first tube in a u.h.f. receiver is most important in raising the signal above the thermal agitation noise of the input circuit, for which reason small u.h.f. types are definitely preferred. Regeneration increases over-all gain without improving the signal-to-noise ratio, provided that increased selectivity in the regenerative stage does not determine the receiver's over-all selectivity.

Superregenerative Receivers. A very effective simple receiver for use at ultra-high frequencies, if properly adjusted, is the superregenerative receiver. The theory of this type

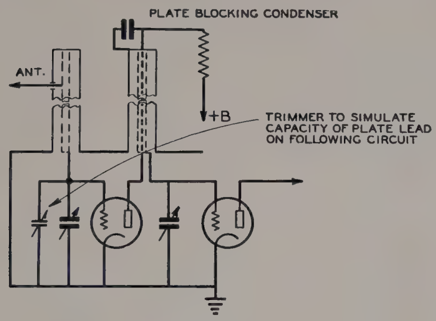


Figure 6.

CONCENTRIC TANK CIRCUITS AS USED IN ULTRA HIGH FREQUENCY RECEIVERS.

Concentric tanks are best at very high frequencies as they have a much higher impedance at these frequencies.

is covered in Chapter 4 and is illustrated in Chapter 18.*

Superheterodyne Receivers. Although they involve the use of more tubes, superheterodyne receivers are somewhat less critical to adjust properly than the superregenerative type. They have the advantages of not causing broad interference locally, and have greater selectivity. The main problem in them is to obtain adequate oscillator voltage injection so that the conversion gain is satisfactory. Screen or suppressor injection requires a strong oscillator if the mixer tube's grid circuit is properly shielded; if it is not, leakage to the control grid will provide grid injection. The latter (often recommended by tube manufacturers for best gain on ultra-high frequencies) results in greatest "pulling" but this can be eliminated by use of a high intermediate frequency and proper construction.

Cathode injection is not recommended by manufacturers because a long cathode lead increases the *transit time effect* and decreases the apparent input resistance of the tube; however, at very high frequencies, several good receivers have used this variation of grid injection by having the mixer cathode clip tap directly on the oscillator tank with very little inductance from the tap to ground and to the grid and plate r.f. return leads.

A stable, hum-free oscillator is necessary in a u.h.f. superheterodyne. Small tubes like the 9002 are satisfactory for this purpose. Heater chokes may reduce hum in cathode-above-ground circuits. Doubler-oscillator circuits or a very high i.f. can be used to reduce the oscillator frequency. Crystal controlled oscillators can be used when the i.f. channel is a tunable receiver.

Here again, an r.f. stage is advantageous to prevent the oscillator from radiating, and to obtain the best signal-to-set-noise ratio, the gain of the r.f. stage being higher than for the mixer, with its output riding over subsequent noise in the receiver. The use of sections of transmission lines instead of coils can improve gain and simplify adjustment and ganging.

High signal input resulting from the use of a carefully designed antenna and feed line, and properly adjusted coupling to the input circuit of the receiver, are essential in obtaining maximum performance. Balanced or shielded feed lines, to reduce pick-up of undesired outside noise, are helpful. The best antenna systems are generally those that are most effective at angles close to the horizontal.

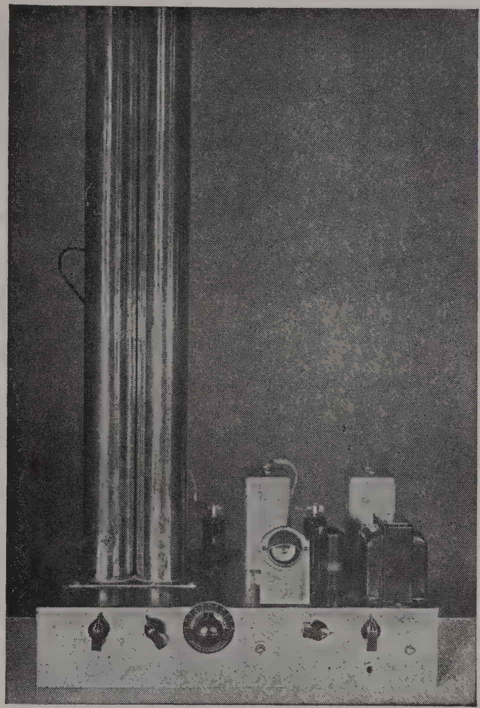


Figure 7.

SUPERHETERODYNE FOR 56 MC. USING CONCENTRIC TANK CIRCUITS.

The acorn tubes used in the high frequency stages are located under the chassis.

*For a more extensive study of its basic theory and adjustment, see articles by Frederick W. Frink in RADIO for March and April, 1938, and a review by E. H. Conklin in RADIO for January, 1941.

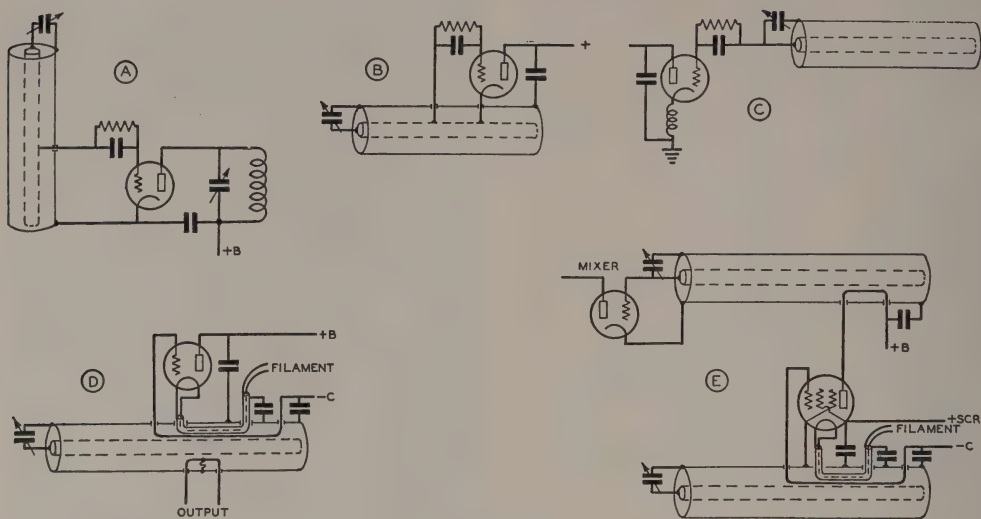


Figure 8.

TYPICAL COAXIAL LINE CONTROLLED OSCILLATOR CIRCUITS.

(A) Concentric line tuned grid, coil tuned plate oscillator. (B) Cathode-above-ground type oscillator circuit with concentric line. (C) Single control oscillator circuit without tap on line, although stability can be increased by tapping the grid down. (D) RCA's oscillator circuit used in a broad band transmitter having good stability, requiring only one tuned circuit. (E) Similar to (D) but showing pentode tube and balanced loop coupling to mixer stage. All coaxial tanks are shorted at the end opposite the tuning condenser.

Transmitter Theory

At ultra-high frequencies, simple but well constructed stabilized oscillators coupled directly to the antenna are satisfactory for c.w. at 28 and 56 Mc., and for modulated waves above 60 Mc. Master oscillators can be built to drive modulated amplifiers with adequate frequency stability. Where highly stable transmission is desired, however, the tendency among amateurs is to use a crystal or electron coupled oscillator at a lower frequency, followed by frequency multipliers. This arrangement provides good stability under modulation, but may drift in frequency more with heating than will a well designed transmission-line-controlled u.h.f. oscillator.

Single-ended oscillator and amplifier stages are often used, but there is reason to prefer push-pull circuits in order to reduce tube capacity across resonant circuits, to obtain balanced arrangements, and to reduce the importance of the cathode leads.

In oscillators, it is highly important to have a lightly loaded, high Q circuit to control the frequency. Such circuits can substantially reduce hum, drift, and frequency modulation. Partial neutralization is a help. A concentric line (when not used with a poor loading condenser) with loose coupling to the grid of the oscillator tube will turn out a good job in a single-ended or push-pull circuit. More commonly, parallel rods are used in push-pull circuits, particularly in plate circuits; if they have a large diameter, remarkably good stability can be obtained.

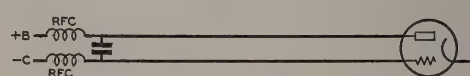


Figure 9.

SIMPLIFIED SCHEMATIC OF SINGLE TUBE OSCILLATOR USING RESONANT LINE WITH PARALLEL CONDUCTORS.

Tubes with an amplification factor of more than 10 are not well suited for use in this circuit. The blocking condenser serves as a shorting bar when frequency adjustment is required. The amount of feedback can be controlled over certain limits by varying the bias resistor or bias voltage.

Due to the appreciable length of cathode leads in terms of wavelength at ultra-high frequencies, push-pull transmitters sometimes become inoperative or unusually inefficient as the frequency is raised. A section of small-size

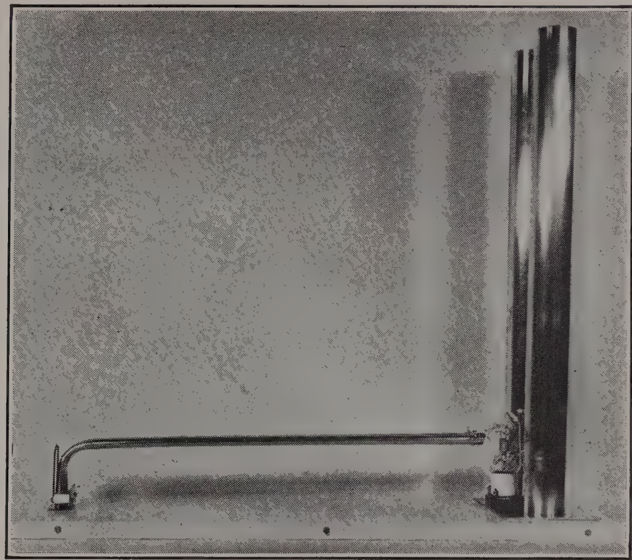


Figure 10.
TYPICAL U.H.F. PUSH-PULL OSCILLATOR USING CLOSE-SPACED RESONANT PIPES FOR FREQUENCY CONTROL.
A "Twin-30" special u.h.f. dual triode is used and permits high efficiency at 224 megacycles.

transmission line electrically a half wavelength long can be used to interconnect filaments and place them at ground potential, as indicated by Figure 13. The shorting bar can be moved to the place where output is greatest or, in some cases, to the only place where oscillation will occur. This application of resonant lines should not be confused with the tuned-plate tuned-grid circuit in which the grid line is moved around to the filament and adjusted to provide the reactance common to grid and plate circuits necessary to maintain oscillation.

Neutralizing condensers are often used on u.h.f. oscillators, being adjusted on either side

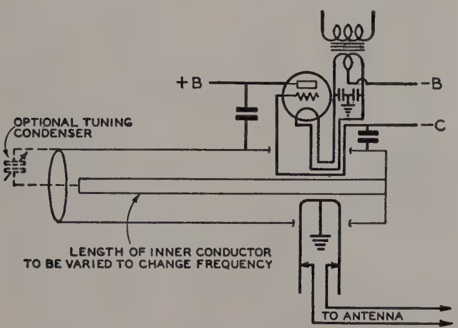


Figure 11.
COAXIAL PIPE OSCILLATOR USING SINGLE TANK CIRCUIT.
The frequency can be varied either by the optional tuning condenser shown or by varying the length of the inner conductor of the concentric line.

of true neutralization, in order to control the amount of feedback and to reduce the effect of tube and plate circuit variations upon the frequency-controlling grid circuit.

Two band operation in oscillators using parallel rods can be arranged conveniently by shorting the open end of the grid control line with a second shorting bar, and readjusting the plate circuit. The resulting half wavelength grid line is loaded by the tube input capacity, making it desirable to slide the grid taps down farther, and requiring a very much shortened line. For instance, a quarter wavelength grid line on 112 Mc. may be 19 or more inches long, whereas a loaded half wavelength line on 224 Mc. may turn out to be only $9\frac{1}{2}$ inches, making it necessary to slide the upper or second shorting bar down from the former open end of the line.

As in the case of receivers, good antennas are helpful, and low angle power is most useful. Less trouble is reported with the proper adjustment of antennas for transmitting than for receiving, however, probably because there is power available with which to work.

Amplifier Hints

The driving power required by an amplifier tube can be high if there are leads of any appreciable length from the grid or plate to any tuning condenser other than one used as a shorting bar on a pair of rods, or if the condenser has a long inductive path through its frame. The returns from these circuits to the cathode are important, especially in single-ended stages. Lead inductance can be reduced

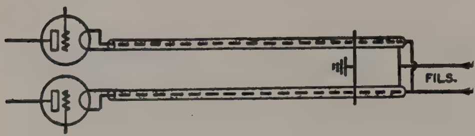


Figure 12.
Arrangement for using shortened ½-wave line in filament circuit to put both filaments at exact ground potential.

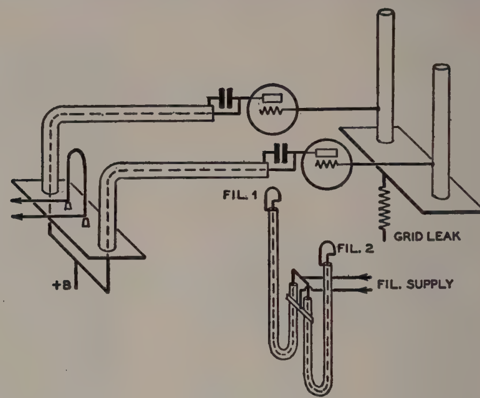


Figure 13.
Practical physical layout for push-pull oscillator using resonant lines in filament, grid, and plate circuits.

by using copper ribbon or tubing for connections, instead of smaller wire.

Frequency doublers have been used to 224 Mc. Push-pull triplers, especially when some regeneration is permitted by using a dual frequency grid circuit or a tuned cathode circuit, are highly satisfactory even above 224 Mc. when suitable tubes are used.

Oscillation difficulties often arise in beam tetrodes due to the resonant frequency of the screen circuit. Where this occurs and cannot be corrected by changing the screen by-pass condenser or its position, a small choke can be inserted in the screen lead before the by-pass condenser.

Both in receivers and transmitters, regeneration or oscillation often results from the use of cathode bias, not adequately by-passed for u.h.f. Ordinary by-pass condensers have considerable inductance in them which combined with their capacity may place a sizable reactance in common with the grid and plate returns. Small silvered mica condensers have sometimes proved better than units of average size and higher capacity. Special u.h.f. sockets with built-in by-pass condensers can be used to advantage above 200 Mc.

Centimeter Waves and Microwaves

With the advent of specially built tubes, it is no longer difficult to obtain appreciable power at ¾ meter (75 centimeters, 400 Mc.) and beyond. The W.E. 316-A will deliver 5 watts or more at 400 Mc., while the RCA 1628 as an amplifier is rated at 50 watts input at 500 Mc. and 43 watts at 675 Mc., the output depending on the circuit and efficiency.

A relatively new development is the velocity-modulated Klystron, with which an output of 100 watts can be obtained at 750 Mc. in an oscillator-amplifier set-up. The tube is like a cathode ray tube, with the stream of electrons passing a hole in a surrounding copper can.

Due to the "cavity resonance" of the chamber, which is essentially a self-enclosed quarter-wave transmission line, power is developed within it which can be delivered to the load by means of a half-turn coupling loop. These tubes were made available under the description "RCA-825 Inductive Output Amplifier." They are designed for use at frequencies of 300 Mc. and *above,* where they are capable of power outputs of 35 watts. A relatively high degree of efficiency is attainable with this type of amplifier stage, 60 per cent efficiency being typical at 500 Mc. Power is placed on the "collector," or plate, which is rated at a maximum of 2000 v.d.c. and 50 ma. The higher voltages required on the other elements are attainable at low-cost, as in cathode-ray tube circuits, because of the insignificant current required. The rated collector dissipation of the tube is 50 watts.

Further U.H.F. Data. For information on transmitters, receivers, and antennas for use on the ultra-high frequencies, turn to Chapters 18, 19, and 22.

CHAPTER EIGHTEEN

U. H. F. Receivers and Transceivers

56 MC. CONVERTER

For receiving stabilized amplitude modulated signals on 56 Mc., an ordinary communications receiver can be used in conjunction with a suitable converter. The converter illustrated in Figures 1 and 2 will be found highly sensitive and ideal for the job.

A high gain mixer using either an 1852 or 1231 receives injection voltage from a 6C5, 6J5, or 7A4 "hot cathode" oscillator.

Construction

The photograph illustrates the layout. A small stock cabinet and the chassis designed for it form the basis for the unit. Mounted in the center of the panel is a small 25-$\mu\mu$fd. per section dual-stator variable. The section nearer the panel, tuning the mixer input, has only one remaining stator plate; the rear portion, for the oscillator, has all but two stator plates removed. This condenser is mounted with the four tapped holes in the frame pointing upward. These holes are then used to support a shield which in addition to covering the condenser also acts as a baffle between the two coils.

Directly back of the tuning gang is the 1852 mixer; to the left is the oscillator coil, and to the right, the mixer coil. The can behind the 1852 contains a tuned output coil and link coupling to the receiver used as an i.f. channel. Below the tuning gang is a 15-$\mu\mu$fd. trimmer on the mixer to eliminate tracking problems on separate bands.

All oscillator leads should be made rigid to avoid shock detuning of the circuit. The ground leads are all brought to one point, which is even more advisable in the mixer circuit where an extra fraction of an inch in the cathode lead, common to both the grid and plate returns, is undesirable in that it affects the gain.

The converter is designed to work into a receiver tuned to a spot between 3000 and 3500 kc. The output coil L_4 is simply a midget b.c.l. antenna coil of the type having a low impedance primary. The coil is tuned by the mica trimmer C_6 and used backwards, the "primary" acting in this case as the secondary.

In some cases, operation will be improved by connecting a .0005-μfd. midget mica condenser directly from the plate of the 6C5 to ground.

Adjustment

The first step in lining up the converter is to adjust the output circuit to resonance with the receiver used as an i.f. amplifier. This is easily done, inasmuch as the receiver noise,

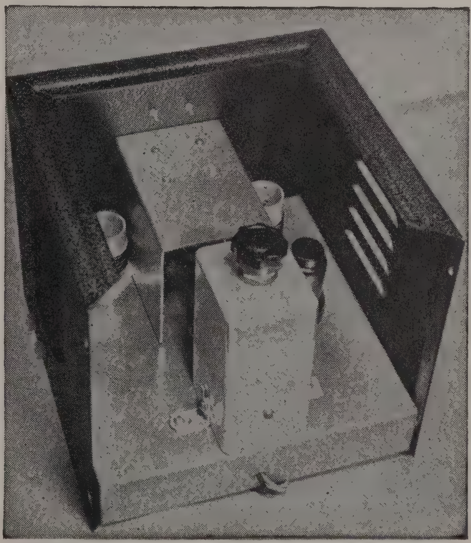


Figure 1.

INSIDE THE 1852 U.H.F. CONVERTER CABINET.

The two-gang tuning condenser is under the U-shaped shield between the two coils. The can in the foreground houses the output coil, L_4, and its trimmer, C_6. Directly behind this can and hidden from view is the 1852; the 6C5 may be seen to the right.

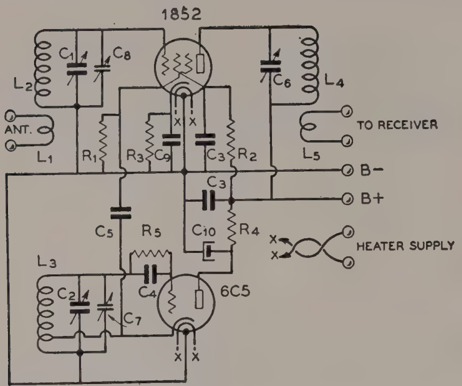


Figure 2.
GENERAL WIRING DIAGRAM OF THE 1852 CONVERTER.

C_1, C_2—Dual 25-$\mu\mu$fd. midget, altered as described in text
C_3—.01 mica
C_4, C_5 — .00005 - μfd. mica
C_6 — 100-$\mu\mu$fd. mica trimmer
C_7—25-$\mu\mu$fd. air trimmer
C_8—17.5-$\mu\mu$fd. midget
C_9—.01-μfd. mica
C_{10} — 8-μfd. 450-volt electrolytic
R_1—25,000 ohms, ½ watt
R_2 — 40,000 ohms, 1 watt
R_3—1500 ohms, 1 watt
R_4—5000 ohms, 1 watt
R_5—50,000 ohms, ½ watt
L_1 — 3 turns at cold end of L_2
L_2 — 3 turns on 1" form spaced dia. of wire
L_3—3¾ turns on 1" form spaced dia. of wire. Cathode tapped ¾ turn from cold end
L_4, L_5—Solenoid type midget b.c.l. antenna coil, half of turns removed from both windings

due both to shot effect in the mixer tube and signal or background racket at the i.f., increases when the circuit is brought in tune. The oscillator can be tuned around to locate a signal, but an easier way to set the oscillator is to listen for it in an all-wave receiver and set it at 28 Mc. plus the i.f.

When this adjustment has been made, there remains only to line up the mixer input circuit on outside noise or on a signal, using the trimmer on the panel (which also acts as a gain control). Ordinarily it will be necessary to obtain proper antenna coupling, inasmuch as high antenna pick-up and transfer to the mixer input will be important in determining weak-signal sensitivity and signal-to-noise ratio.

Voltage Regulation. If plate voltage fluctuations are sufficient to cause an objectionable shift in the oscillator frequency, as might be the case with an a.c. power pack running from a line to which several large intermittent loads are connected, the oscillator plate voltage can be stabilized simply by hooking a VR-150-30 type voltage regulator tube between the low side of R_4 and ground. The VR tube should be shunted by a .05-μfd. tubular condenser. The plate supply should have at least 225 volts for the VR tube to function properly.

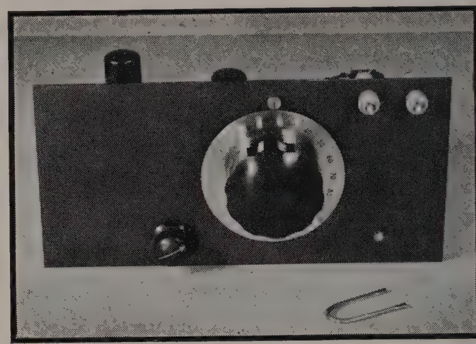


Figure 3.
SIMPLE RECEIVER FOR USE BETWEEN 112 AND 350 MC.
The set is a superregenerator with a 9002 oscillator and plug-in inductances. The "coil" in the foreground is for 224 Mc.

SUPERREGENERATIVE RECEIVER FOR 2½-, 1¼-, AND 1-METER OPERATION

The superregenerative receiver illustrated in Figures 3–6 can be used on the 112, 224, or "over 300" Mc. bands by means of plug-in inductances. With the shortest possible jumper of no. 8 wire plugged into the coil socket, the frequency is about 350 Mc.

The oscillator utilizes a 9002 midget triode, which electrically is the same as the acorn prototype 955 but is cheaper in price and uses a more conventional socket. The important feature of the oscillator is the use of the smallest possible components and shortest possible leads, with nothing but Isolantite or polystyrene insulation comprising or touching the r.f. components.

To obtain the shortest possible leads, the tube is mounted on its side as illustrated in Figure 5. An Amphenol polystyrene crystal holder socket is used as a coil jack. The tuning condenser is the smallest size air trimmer with all but one stator and one rotor plate removed, the two plates being double spaced. The shaft is driven through a ceramic insulated coupling.

No regeneration control is provided, the antenna coupling being increased to the greatest value which will permit superregeneration.

Considerable feeder loss will be present on 224 Mc. and above, even with the best trans-

mission line. Therefore, the transmission line should be as short as possible, besides being of good quality.

The $2\frac{1}{2}$-meter coil consists of 6 turns of no. 14 enamelled wire on a $\frac{1}{2}$-inch inside diameter, spaced to $\frac{5}{8}$ inch. The ends are sweated into "phone tip" plugs or pins from the base of a discarded tube having less than 8 prongs.

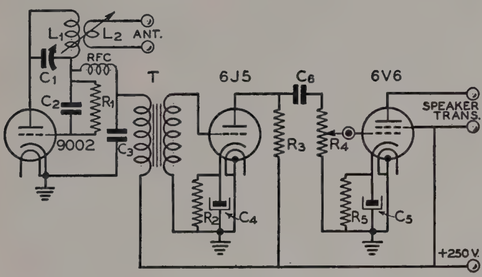
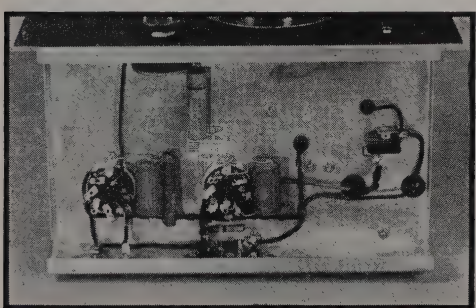


Figure 4.
WIRING DIAGRAM OF U.H.F. SUPERREGENERATOR.

C_1—Small "trimmer" type 5-plate midget variable air condenser with all but one rotor and one stator plate removed, the two remaining plates being double spaced
C_2—Smallest size 50-$\mu\mu$fd. mica fixed condenser
C_3 — .006-μfd. mica
C_4, C_5—10-μfd. 25 v. electrolytics
C_6 — .05-μfd. tubular paper
R_1—500,000 ohms, ½ watt
R_2 — 2000 ohms, ½ watt
R_3 — 50,000 ohms, ½ watt
R_4—100,000-ohm pot., a.f. taper
R_5 — 400 ohms, 1½ watts
L_1—Refer to text
L_2—1 or 2 turns hookup wire coupled to L_1, using fewest no. of turns which will permit sufficient coupling
T — Midget 1-3 ratio interstage transformer

Figure 6.
UNDER-CHASSIS VIEW OF U.H.F. SUPERREGENERATOR.
Placement and wiring of the a.f. and d.c. components is not particularly critical.



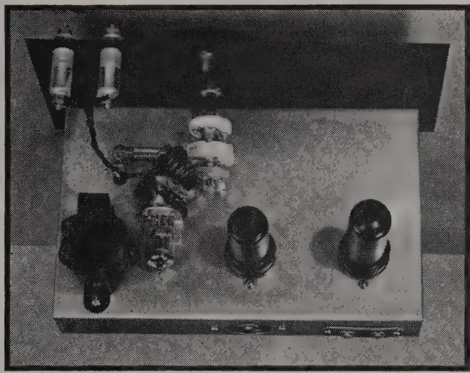


Figure 5.
CONSTRUCTION DETAIL OF U.H.F. SUPERREGENERATOR.
All r.f. components are placed as close together as possible. All insulation is either polystyrene or Isolantite.

The $1\frac{1}{4}$-meter coil consists of a $4\frac{1}{2}$-inch length of no. 8 bare copper wire bent into the shape of a "U." For higher frequencies, up to approximately 350 Mc., a correspondingly shorter "U" is required.

The frequency of the $2\frac{1}{2}$-meter coil may be adjusted to hit the band by spreading or compressing the turns. The higher frequency "coils" can be similarly adjusted (but over narrower limits) by shoving them in or out of the coil socket a short distance.

The last filter condenser in the power pack feeding the receiver should have a capacity of at least 8 μfd. or else there may be a tendency for the receiver to "motorboat."

It is inadvisable to permit the plate voltage to be applied to the 9002 for more than a few seconds when it is not oscillating, as the high value of plate current drawn by the tube in a non-oscillating condition is sufficient to damage the tube if allowed to persist.

The two feed-through type antenna terminal insulators are satisfactory for a close-spaced, 2-wire open line. Such a line is less expensive than a good low loss coaxial line, but will radiate considerably on $1\frac{1}{4}$ meters. If a coaxial line is to be used, a coaxial type terminal connector should be substituted.

112-MC. SUPERHET FOR EITHER AMPLITUDE OR FREQUENCY MODULATION

The 112-Mc. superheterodyne illustrated in Figures 7–10 provides excellent performance on either amplitude modulated (a.m.) or frequency modulated (f.m.) signals. The i.f.

channel is broad enough that amplitude modulated oscillators can be received satisfactorily if the oscillator is reasonably stable.

High sensitivity is provided by the use of an acorn pentode and a coaxial pipe tank circuit in the mixer, in conjunction with control grid injection of the oscillator voltage.

For reception of amplitude modulated signals, the limiter is "opened up" by means of switch S_1 (which is operated by turning R_{17} full off) and the 6H6 discriminator is changed to a diode demodulator by means of switch S_2.

Construction

The chassis, which is surmounted by an 8 x 17-inch panel, measures 7 x 15 x 3 inches. The 956 is located near the left rear corner of the chassis, with its concentric grid tank running along the rear of the chassis, as is apparent from the photographs. The concentric tank is held to the chassis by two copper straps, one near each end. The mixer grid condenser is placed between the 956 and the left edge of the chassis, making it convenient to secure short leads to both the mixer and the inner conductor of the tank circuit.

To help in obtaining short leads, the oscillator socket has been mounted with its base above the chassis, making it necessary that the 6J5GT be located under the chassis. The oscillator grid coil is supported from the tuning condenser on one end and the no. 1 socket terminal on the other. The plate by-pass, C_{24}, is located right at the socket and connected in the shortest possible manner between the plate and no. 1 terminal. A dial having a built-in planetary reduction unit is used on the oscillator to allow accurate tuning.

To aid in isolating the oscillator and mixer from each other so that the injection may be controlled by pushing the lead from the mixer grid in and out of the outside conductor of the mixer tank circuit, a 3 x 4-inch copper shield is placed between the two stages. The shielding is supported by small angle brackets.

The first i.f. transformer, T_1, is located directly in front of the mixer, with the first 1852 between this transformer and the panel. The second i.f. stage with its associated transformers, T_2 and T_3, runs along the front of the chassis from left to right. Behind T_3 is the 6SJ7 limiter, which feeds through the discriminator transformer at its right to the 6H6 discriminator between the transformer and the panel. The audio follows along the right edge of the chassis, while the VR-150 regulator is located behind T_4.

The only wiring precaution that needs to be observed is keeping the grid and plate leads short. This holds for the i.f. section as well as for the high frequency circuit. No regenera-

Figure 7.
LOOKING DOWN FROM THE FRONT OF THE 112-MC. F.M.-A.M. RECEIVER.
The adjustable coupling lead from the oscillator grid through the concentric mixer grid tank is visible in this photograph. The controls are, from left to right, mixer tuning, oscillator tuning, limiter "threshold" and limiter cut out, audio gain, and f.m.-a.m. switch.

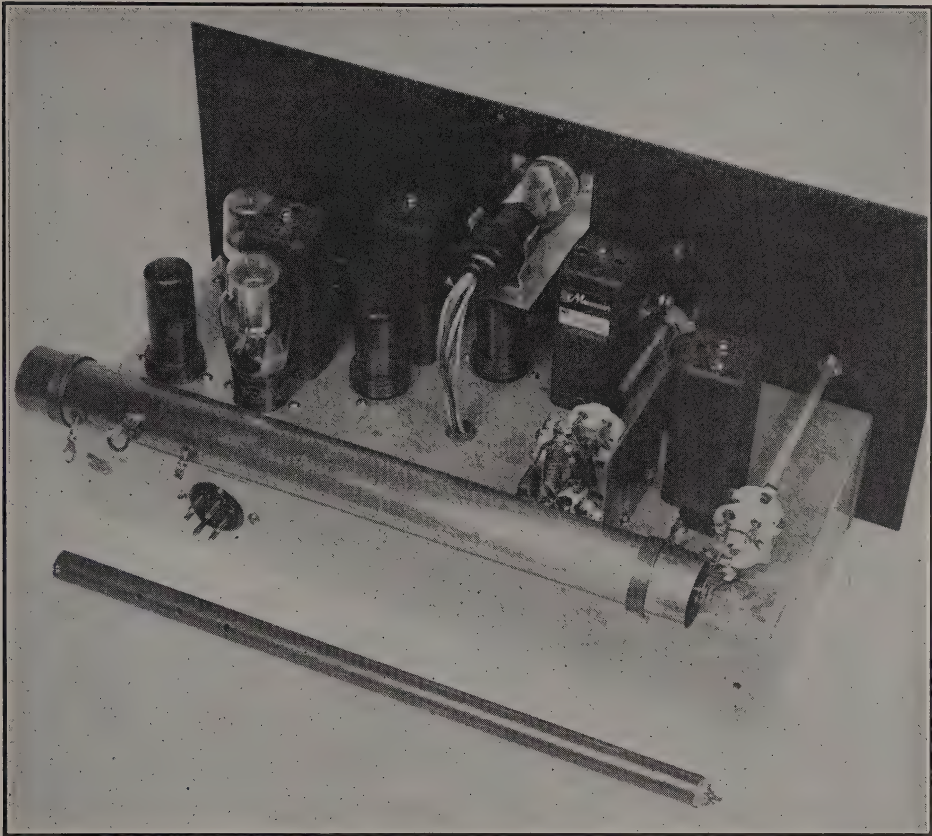


Figure 8.
CONSTRUCTION DETAILS OF THE 112-MC. F.M.-A.M. RECEIVER.
The outer conductor of the concentric pipe tank is held firmly and grounded electrically to the chassis by means of a narrow copper strap at each end of the tank. The shield partition between the oscillator and mixer circuits is necessary for good stability. The smaller diameter tank shown in the foreground works almost as well as the large one, and may be substituted if desired. The inner conductor of the smaller tank is held in position at the unshorted end by means of a polystyrene spacer.

tion trouble in the i.f. section should be experienced if the grid and plate leads run directly from small holes below the i.f. transformer to their proper terminating point on the sockets.

The mica by-pass and coupling condensers in the mixer and oscillator sections should be of the smallest physical size available, since a physically small .00005-μfd. condenser will often prove to be a better by-pass or coupling device at 112 Mc. than a .002-μfd. or larger mica condenser having proportionately larger dimensions.

The Coaxial Tank. The mixer tank consists of a 14-inch length of $1\frac{3}{8}$-inch copper pipe as the outer conductor and a $\frac{3}{16}$-inch copper tubing inner conductor. These conductors give a radius ratio of approximately 7–1, which seems to be a good compromise between impedance, Q, and overall tank size.

No actual "shorting disc" is used with the line shown in the receiver. The inner conductor is merely flattened at the "closed" end of the tank and two short right-angle bends made to allow it to be held to the outer conductor with a screw. This method is perfectly permissible where extremely high Q in the line is not necessary.

The antenna coupling "loop" is a piece of no. 10 wire covered with "spaghetti" where it is inside the tank, and supported within the tank by being run through tight fitting grom-

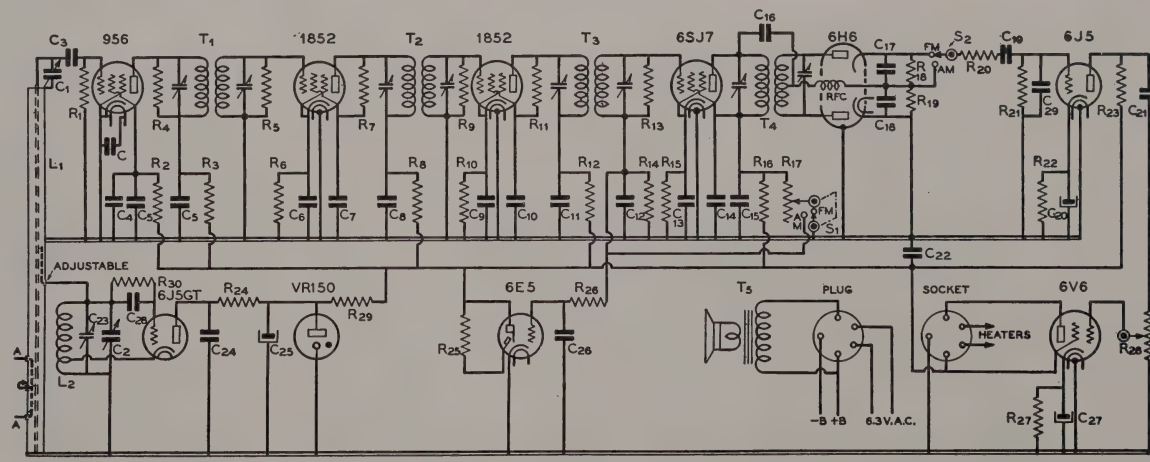


Figure 9.
WIRING DIAGRAM
OF THE
F.M.-A.M.
RECEIVER.

VALUES OF COMPONENTS

C—.0001-μfd. mica
C_1—7-μμfd. midget with one stator plate removed
C_2—15-μμfd. midget variable
C_3, C_4—.0001-μfd. mica
C_5, C_6, C_7, C_8, C_9, C_{10}, C_{11}—.01-μfd. 600-volt tubular
C_{12} — .0001 - μfd. mica
C_{13}, C_{14}, C_{15}—.01-μfd. 600-volt tubular
C_{16} — .00005 - μfd. mica
C_{17}, C_{18}—.0001-μfd. mica
C_{19} — .01-μfd. 600-volt tubular
C_{20} — 10 - μfd. 25-volt electrolytic
C_{21} — .01-μfd. 600-volt tubular
C_{22} — 0.1-μfd. 600-volt tubular
C_{23} — 2-35 - μμfd. mica trimmer
C_{24} — .0005 - μfd. mica
C_{25} — 8-μfd. 450-volt electrolytic
C_{26} — .01-μfd. 600-volt tubular
C_{27} — 10-μfd. 25-volt electrolytic
C_{28} — .0001 - μfd. mica
C_{29} — .0005 - μfd. mica
R_1—5 megohms, ½ watt
R_2—100,000 ohms, 1 watt
R_3—2000 ohms, ½ watt
R_4, R_5 — 50,000 ohms, ½ watt
R_6—150 ohms, ½ watt
R_7 — 30,000 ohms, 1 watt
R_8—2000 ohms, ½ watt
R_9 — 50,000 ohms, ½ watt
R_{10}—150 ohms, ½ watt
R_{11}—30,000 ohms, 1 watt
R_{12} — 2000 ohms, ½ watt
R_{13}—50,000 ohms, ½ watt
R_{14}—250,000 ohms, ½ watt
R_{15}—100 ohms, ½ watt
R_{16}—75,000 ohms, 1 watt
R_{17} — 10,000-ohm wire-wound potentiometer
R_{18}, R_{19} — 100,000 ohms, ½ watt
R_{20}—50,000 ohms, ½ watt
R_{21}—500,000 ohms, ½ watt
R_{22}—2000 ohms, ½ watt
R_{23}—50,000 ohms, 1 watt
R_{24}—3000 ohms, 1 watt
R_{25}—1 megohm, ½ watt (supplied with 6E5 socket assembly)
R_{26}—1 megohm, ½ watt
R_{27}—500 ohms, 10 watts
R_{28} — 500,000-ohm potentiometer
R_{29}—3000 ohms, 10 watts
R_{30}—100,000 ohms, ½ watt
T_1, T_2, T_3, T_4 — 3000 kc. output i.f. transformer. See text for alterations to T_4
T_5—Pentode-plate-to-voice coil transformer (on speaker)
S_1—S.p.d.t. switch (on R_{17})
S_2—S.p.d.t. toggle switch
RFC—2½ mhy.
L_1 — 14″ copper concentric line. Outer conductor 1⅜″ o.d., inner conductor 3/16″ o.d. See text
L_2—5 turns of no. 16 bare copper, ¼″ inside diameter and wound to a length of ½″. Cathode tap 1½ turns from ground end

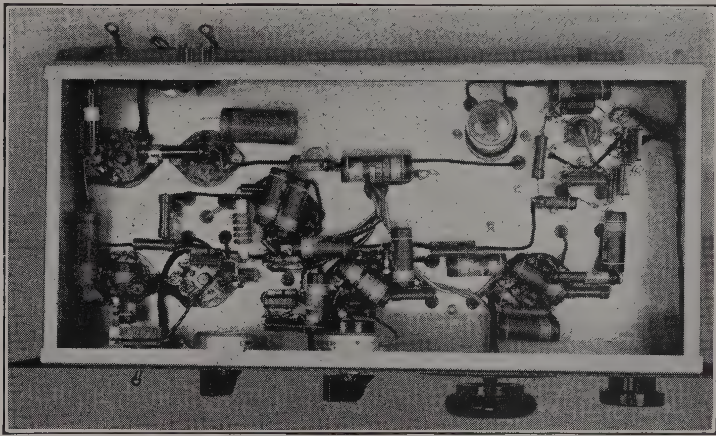


Figure 10.
UNDER-CHASSIS VIEW.
The 956 mixer and the 6J5GT oscillator may be seen under the chassis of the receiver. The two lugs protruding from the concentric tank at the upper left of the photograph are for antenna connections.

mets in the outer conductor. A lead soldered to the center of the loop inside the tank is brought out and provided with a lug to enable the center of the loop to be grounded when a balanced, 2-wire feeder is used. The end of the loop nearest the shorted end of the tank is grounded when a single-feeder type antenna is used. The loop is $2\frac{1}{2}$ inches wide, but experiment will probably be necessary to obtain optimum coupling with lines of different impedance than the 400-ohm feeder used with the original receiver. Coupling adjustments are made by pushing the loop toward or away from the inner conductor.

If desired, a smaller diameter tank may be used, so long as the outer conductor is at least $\frac{1}{2}$ inch in diameter and the conductor ratio is kept between 6 and 10. Unless the ratio is exactly 7, the length of the tank will have to be altered slightly. The performance will be practically as good as with the $1\frac{3}{8}$ inch diameter tank.

The Discriminator Transformer. As received from the manufacturer the transformer, T_4, specified in the diagram has no center tap on its secondary and lacks sufficient coupling to serve as a discriminator transformer. Consequently, the transformer must be altered as follows: After removing the transformer from its shield can, the lower winding, which is to become the secondary, is completely unwound from the dowel. If the unwinding is done carefully, a narrow ridge of the compound with which the windings are impregnated will be left on each side of the space the winding occupied. These ridges will form a sort of "slot" in which to rewind the wire which has been removed. It will be found that about 65 turns of wire were on the winding, but it will be impossible to get more than 55 to 58 turns back in the slot by hand scramble-winding methods. In the receiver shown, a trial rewinding of the wire indicated that 56 turns could be replaced, necessitating that the center tap be brought out at the 28th turn.

After the secondary has been rewound on the dowel, it should be thoroughly covered with Duco cement or a similar coil compound, and allowed to dry for an hour or more. When the cement has dried thoroughly, it will be found that a firm pressure against the winding will allow it to be slid along the dowel toward the primary to increase the coupling between the windings. The proper location for the secondary is a position where the distance between the adjacent edges of primary and secondary is about $\frac{1}{8}$ inch. Another coating of the coil dope will hold the winding in place, and the transformer may be reassembled in its shield can and installed in the receiver.

Adjustment

Aligning the I.F. Channel. There is no really simple way of accurately aligning the i.f. and discriminator in an f.m. receiver. The inclusion of a 6E5 "magic eye" tube operating from the voltage developed across the limiter bias does help considerably, however, aside from its intended use as an accurate tuning indicator for placing f.m. signals "on the nose." Probably the easiest method of aligning the receiver is first to couple loosely an ordinary tone-modulated signal generator to the plate of the mixer stage. With both switches set for "a.m." make a rough alignment for maximum audio output. This assumes that the i.f. transformers are somewhere in the vicinity of alignment so that some sort of signal may be forced through the i.f. channel to get a start on the trimming process. If no signal is heard at the output

when the signal generator is applied at the mixer plate and tuned around over a narrow range around 3000 kc., it must be assumed that the i.f. transformers are considerably out of alignment and the usual procedure of first coupling the signal generator to the primary of the last i.f. transformer (T_4) and then working back toward the mixer stage must be followed.

After a rough setting of the trimmers has been made the alignment may take on a more exact nature. With the signal generator still applied to the primary of T_1, but with switch S_1 changed to the "f.m." position by cutting in all of R_{17}, each trimmer on the first three i.f. transformers should be adjusted for maximum voltage across R_{14}, as indicated by the closing of the "eye." Next, the setting of the trimmer across the secondary of T_4 should be tackled—and here is where the trimming becomes critical. Since the trimmer adjusting screw is "hot" for r.f., the tool used for this adjustment should be of the low-capacity type having a long composition or wood handle.

The discriminator output switch, S_2, should be thrown to the "f.m." position and—assuming that the primary of T_4 has been set up somewhere near resonance in the previous rough alignment—tuning the secondary winding through resonance should give a very sharp and definite drop in the audio output, the audio-tone volume increasing on either side of resonance but dropping to a very low value or disappearing entirely at exact resonance. The signal from the signal generator should be kept at i.f. resonance, as indicated by the 6E5, during the alignment.

The last adjustment to be made should be that on the primary of T_4. There are two ways of getting this circuit properly tuned. Probably the simplest method is to keep the signal generator tuned right in the "notch" of the secondary winding but increase the amount of signal applied to the i.f. channel until a small amount of audio comes through at this frequency and then tune the primary winding for maximum decrease or "dip" in the remaining audio.

The other method of trimming the primary involves rocking the signal generator back and forth across the resonant frequency previously obtained, observing the strength of the peaks in audio output which are heard on each side of the "notch." When the primary is properly tuned, these peaks will be symmetrically located, one on each side of the "notch" frequency, and of equal strength. If the i.f. loading resistors are of the values indicated under the diagram, and the coupling between the primary and secondary of T_4 has been properly adjusted, the peaks will be approximately 130 kc. apart.

Those who find it more convenient to use an unmodulated signal at the i.f. frequency and a vacuum-tube voltmeter or zero-center high-resistance voltmeter to align the i.f. and discriminator may do so by connecting the indicating instrument between the top of R_{18} and ground and, after aligning the i.f. transformers up to T_3 by the 6E5, adjusting T_4 so that zero voltage is obtained at the center of the i.f. band, and equal and oppositely-polarized voltages are obtained for equal and opposite shifts in signal-generator frequency from center frequency. When a vacuum-tube voltmeter is used for this adjustment it will be necessary to place a battery in series with the instrument to bring it somewhere near half scale.

R.F. Alignment. There is little that need be said about tuning up the front end of the receiver, since the only problem is to find the band. The simplest way to do this is to hunt for a 2½-meter signal with the oscillator padding condenser, C_{23}, keeping the mixer grid aligned by following with C_1. In the absence of signals, the best procedure would be to set the oscillator tuning condenser at mid-scale and adjust the padding condenser so that the oscillator is on a frequency 3000 kc. lower than the center of the band, or 111 Mc. The frequency should be measured by Lecher wires, the proper distance between points being very close to 53 inches. A detailed discussion of the use of Lecher wires is given in Chapter 17. The glow in the VR-150 makes a fairly good resonance indicator for this purpose.

Lining up the mixer grid involves only tuning the mixer grid condenser and adjusting the antenna and oscillator coupling for maximum background or signal. The two coupling adjustments will be found to be somewhat interdependent, and should be adjusted simultaneously. The mixer coupling is not extremely critical, however, and optimum results should be obtained over a wide range of injection voltage. Two inches of wire available for pushing through the grommet and into the mixer grid tank will provide sufficiently wide range of coupling from the oscillator. Too little coupling will result in a loss of sensitivity, while too much coupling will cause bad pulling of the oscillator by the mixer tuning. Fortunately, maximum sensitivity is realized with quite a bit less coupling than is required to cause serious pulling.

56 Mc. Operation. This receiver makes an excellent 56 Mc. f.m. superheterodyne if a suitable coil is substituted for the coaxial mixer tank and a larger coil is substituted for the high frequency oscillator tank. No other changes need be made.

400-MC. SUPERREGENERATOR OR TRANSCEIVER

Illustrated in Figures 11–17 is a highly effective 400-Mc. superregenerative receiver of a design by J. C. Reed. Losses in the transmission line are minimized by making it as efficient as possible and limiting its length. The antenna, coaxial feeder, and oscillator are constructed as an integral unit. The a.f. section is conventional, and is housed in a separate unit. It may be patterned after that of Figure 4, for use only as a receiver, or after Figure 24 if it is to be used as a transceiver.

For fixed station use, the unit may be mounted outdoors with the oscillator proper in a weatherproof "dog house" and tuned remotely either by means of a small, reversible "tuning motor" such as that employed on some broadcast sets, or else by means of cord and pulleys. If such an arrangement is impractical, the oscillator may be placed in the operating room and the outdoor antenna fed by means of "u.h.f. type" coaxial cable of about 30 ohms. However, the latter arrangement will not be as efficient because of the losses on even the best line at this very high frequency.

For mobile use in an auto, the oscillator may be mounted on a small projecting shelf fastened by means of removable clamps to the car window sill. This will put the radiating portion of the antenna above the top of the car on most of the later models.

The oscillator is of the cathode-above-ground type using a concentric line. This concentric line is made of a 2-inch copper pipe $4\frac{1}{2}$ inches long for the outside conductor, and a $\frac{1}{2}$-inch copper pipe of the same length for the inside conductor.

The diameter of the inner conductor is not critical. For maximum Q the ratio of conductors should be about 3.6, although a much higher value can be used without a noticeable effect.

The cathode connection and one filament connection are tied together by a copper bar soldered directly to the inner conductor 1 inch from the end. Grooves are filed in the bar for the cathode and filament leads in such a way that the tube is held in place by a plate held over the two leads, and thereby acting as a support for the tube. A copper strip is connected to the other filament lead and run down alongside the inner conductor to the cold end

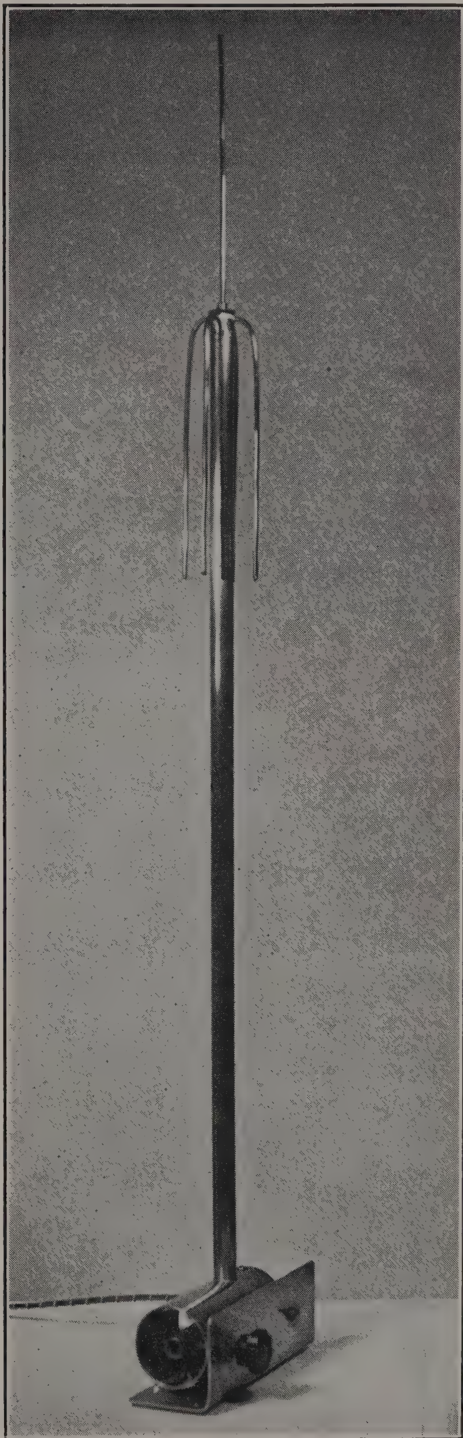


Figure 11.
400-MEGACYCLE OSCILLATOR.
This oscillator may be used either as a superregenerative receiver or as a transceiver. An acorn type 955 triode is used.

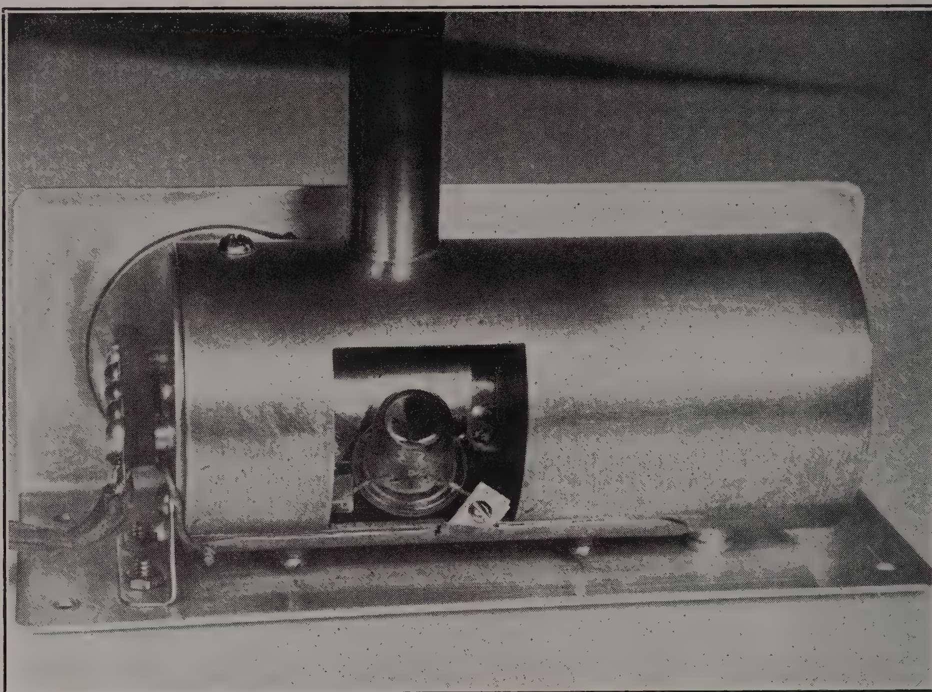


Figure 12.
CONSTRUCTION OF MICROWAVE "PIPE" OSCILLATOR.
A portion of the outside pipe is cut away to permit installation of the tube. Observe the "built in" by-pass condensers.

of the pipe. This filament lead is insulated from the inner conductor by a thin sheet of mica. Equal results can be obtained with the filament lead down through the inside of the inner conductor by drilling a hole in the ½-inch pipe near the cathode connection. If this type of connection is used, the filament wire will have to be by-passed to ground at the cold end of the pipe.

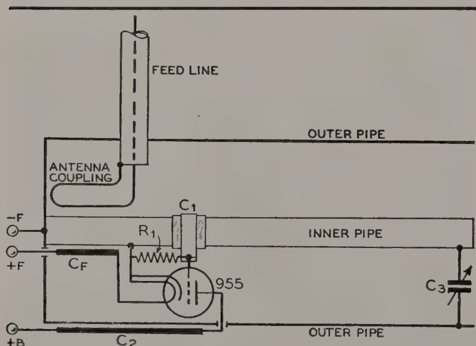


Figure 13.
SCHEMATIC DIAGRAM OF MICROWAVE RECEIVER.
R_1 has a value of 2 megohms. Condensers C_1, C_2, and C_F are described in the text. Tuning condenser C_3 is shown in detail in Figure 15.

The grid is connected approximately half way up the pipe through a mica condenser consisting of a copper band 3/32 inch wide, insulated by a thin sheet of mica. The band is a tight fit over the mica and the pipe. A 2-megohm grid resistor is connected from the grid condenser to the inner pipe.

The plate by-pass to ground is a copper plate 3 x 1¼ inches, and is insulated from the outer conductor by mica.

Tuning is accomplished by varying the distance of a copper sheet 1½ inches wide alongside the end section of the inner conductor. This sheet is of hard-drawn copper so that it will spring back on its own accord, thereby simplifying tuning arrangements. A string belt is run around the tuning shaft to a pulley 1¾ inches in diameter to act as a tuning indicator. With this size tuning condenser the receiver will tune approximately 75 Mc. each side of the band. This overlap was considered

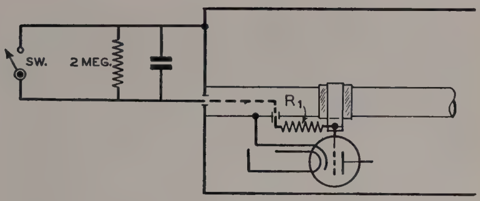


Figure 14.
BASIC MODIFICATION FOR TRANSCEIVER USE.
Resistor R_1 is chosen so as to make the tube draw normal plate current as a transmitter. It normally will be between 5000 and 20,000 ohms, ½ watt. Plate voltage when transmitting should not exceed about 150 volts.

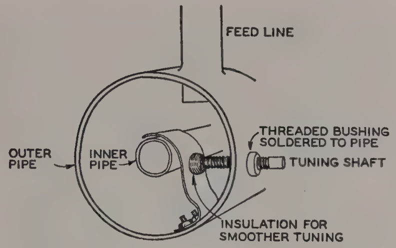


Figure 15.
TUNING ARRANGEMENT.
The tuning shaft is turned by means of a small knob, and a cable around the shaft drives an indicator which is provided with a large pulley as described in the text.

desirable because of the narrow width of the 400-Mc. band and because all frequencies above 300 Mc. are open for experimental use.

In mounting the attached concentric feed line for the antenna, a ½-inch hole is cut in the 2-inch pipe near the closed end so that the ½-inch pipe can be soldered directly to the oscillator. An alcohol torch or very heavy soldering copper will be required. The inner conductor of the concentric feed line is connected to a ¼ x 1½-inch copper strip. The other end of this strip is connected to the 2-inch pipe in such a manner that the antenna coupling can be varied by sliding the feed line closer to or farther from the inner conductor of the oscillator.

If a conventional type coaxial line is to be used, a regular terminal connector can be mounted on the outside pipe conductor about 1½ inches up from the bottom, and the hairpin coupling loop arranged the same as for the integral feed line.

If the oscillator is placed some distance from the rest of the receiver, the by-pass condenser for the quench frequency component (usually .002 to .006 μfd.) should be placed

Figure 16.
DETAIL OF RADIATOR AND FEED LINE.
The inner conductor is kept in place by means of three or four polystyrene centering washers cemented in place with liquid polystyrene.

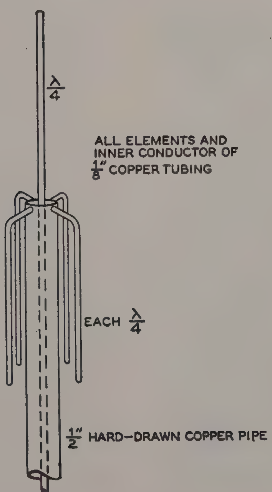
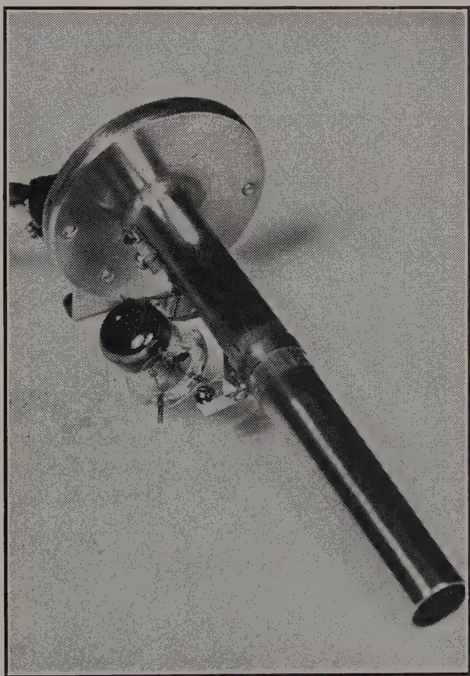


Figure 17.
CONSTRUCTION DETAIL OF OSCILLATOR.
The assembly shown below can be slid out intact after the plate connection to the 955 is disconnected.



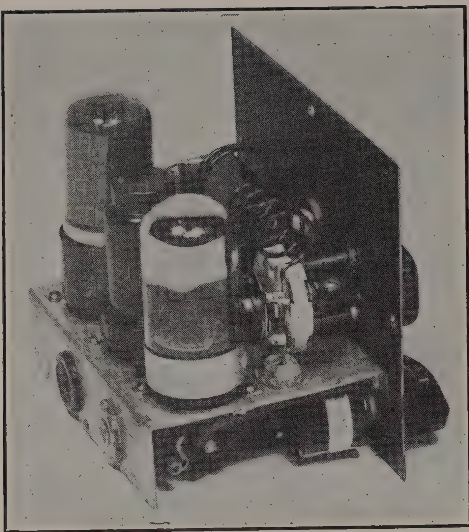


Figure 18.
INTERIOR VIEW OF THE RECEIVER.
The tuning condenser is supported from the front panel by means of two long bolts. The variable antenna coupling coil may be seen in back of the tank coil.

directly across the terminals of the oscillator marked "B plus" and "F minus." If this is not done, radiation from the wire carrying the quench component may cause bad hash in neighboring broadcast receivers.

In Figure 14 is shown the basic modification for using the oscillator in a transceiver arrangement. The resistor R_1 ordinarily will be from 5000 to 20,000 ohms, ½ watt. SW is one section of the transceiver send-receive switch. (See Figure 24.) The midget mica by-pass condenser across the 2-megohm resistor may be from about 10 to 50 $\mu\mu$fd., the exact value having some effect upon the quenching action of the receiver. It should be connected to the bias wire and to the outer conductor end plate with the shortest possible leads right at the point where the wire leaves the end plate.

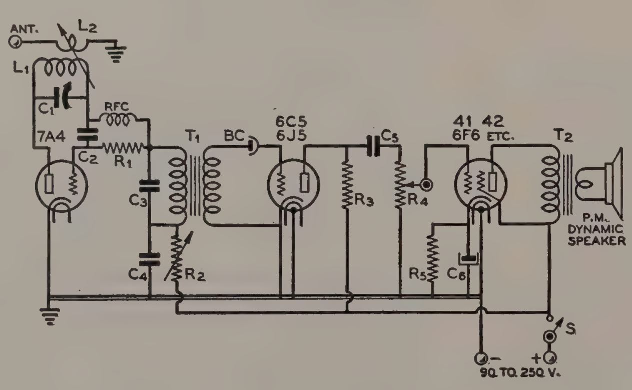
COMPACT 112-MC. SUPER-REGENERATIVE RECEIVER

Illustrated in Figures 18 and 19 is a compact and inexpensive 112-Mc. superregenerative receiver that will give excellent results on amplitude modulated signals either for mobile or fixed station use. It will also work fairly well on frequency modulated signals, especially if the deviation (frequency swing) is comparatively large, but should be considered primarily as an amplitude modulation receiver.

Figure 18 illustrates the arrangement of components. If desired, the receiver need not be made quite so compact; this will simplify the wiring job somewhat.

It is important that a polystyrene or low loss (mica filled) bakelite loktal socket be used for the 7A4 for best results. Also, care should be taken to see that the rotor of the tuning

Figure 19.
CIRCUIT FOR THE SUPERREGENERATIVE 3-TUBE RECEIVER.



C_1—7-$\mu\mu$fd. sub-midget variable condenser
C_2—100-$\mu\mu$fd. midget mica
C_3—.01-μfd. tubular, 400 v.
C_4—0.25-μfd. tubular, 400 v.
C_5—.05-μfd. tubular, 400 v.
C_6—10-μfd. electrolytic, 25 v.
R_1—500,000 ohms, ½ watt insulated
R_2—250,000-ohm midget pot. or a 5000-ohm 1-watt fixed resistor (see text)
R_3—50,000 ohms, 1 watt insulated
R_4—100,000-ohm midget pot., a.f. taper
R_5—600 ohms, 5 watts insulated
T_1—1-3 ratio a.f. transformer
BC—Bias cell, outside shell to grid
L_1, L_2—See coil data
RFC—U.h.f. choke
T_2—Output transformer (on p.m. speaker)

condenser goes to the grid and the stator to the plate. A bakelite or hard rubber shaft extension must be used with the tuning condenser in order to prevent body capacity detuning effects. As an alternative, an insulated coupler may be used in conjunction with a short piece of metal shafting and a panel bearing. Both r.f. choke and grid leak should be connected with the shortest possible leads to the r.f. circuit.

The tank coil, which is soldered directly to the tuning condenser terminals, consists of 4 turns of no. 14 enameled wire, ½ inch in diameter, spaced and trimmed as necessary to hit the band (as determined by Lecher wires).

One of the features of the receiver that results in vastly increased performance and easier tuning is variable antenna coupling. This control has been found of greater importance than the regeneration control, as the latter may be set and left alone if variable antenna coupling is provided. In fact, the regeneration control may be omitted, if desired, in which case a 5000-ohm 1-watt resistor is substituted for R_2.

The antenna coil consists of 2 turns of wire 1 inch in diameter, supported at the grid end of the tank coil. These are cemented with Amphenol 912 to a piece of Lucite or polystyrene ¼-inch shafting, which is supported from the front panel by a pinch-fit shaft bearing. The bearing is placed slightly below the level of the bottom edge of the tank coil in order to permit sufficient variation in coupling. Flexible, insulated wire is used for making connection to the 2-turn antenna coil.

When tuning the receiver, the tightest antenna coupling which will permit superregeneration should be used.

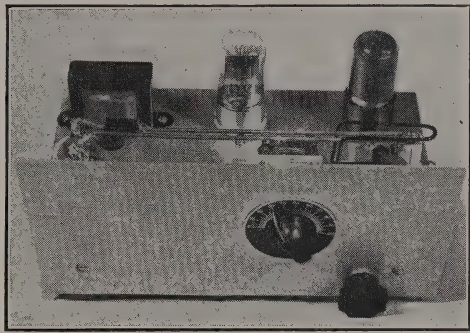


Figure 20.
224-MC. SUPERREGENERATIVE RECEIVER.
An HY-615 triode oscillator and linear tank circuit provide high sensitivity.

224-MC. SUPERREGENERATIVE RECEIVER

Except for the substitution of a linear tank circuit and an oscillator tube better adapted for use at the higher frequency, the 224-Mc. receiver of Figures 20–23 is substantially the same from an electrical standpoint as the 112-Mc. superregenerative receiver of Figures 18 and 19. The mechanical construction is somewhat different, however, as may be seen from Figures 20, 21, and 23.

The receiver is constructed on a 5½ x 11-inch chassis, 1½ inches high, which supports a 5x9-inch front panel. The HY-615 oscillator tube is placed at one end of the chassis, as illustrated, in order to permit horizontal mounting of the linear tank circuit. This tank circuit consists of a length of no. 10 bare copper wire,

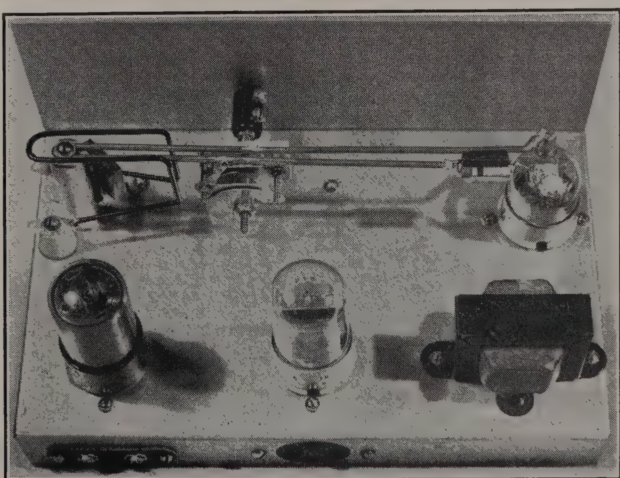


Figure 21.
ILLUSTRATING MECHANICAL CONSTRUCTION OF 224-MC. RECEIVER.
Note particularly the modified tuning condenser and the arrangement of the linear tank and the antenna coupling "hairpin loop."

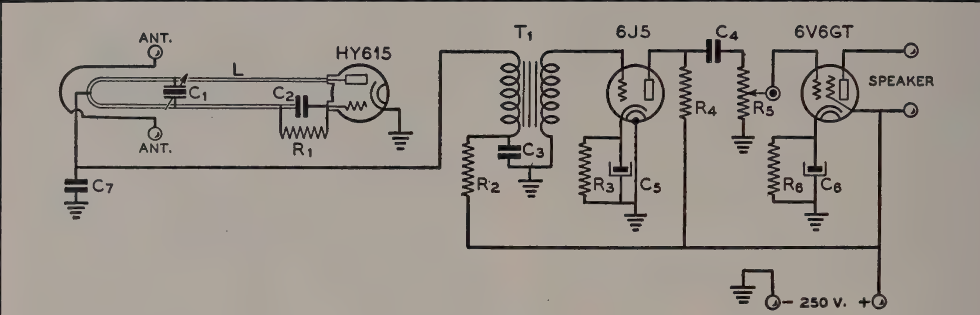


Figure 22.
SCHEMATIC DIAGRAM OF 224-MC. RECEIVER.

C_1—Modified midget condenser, see text
C_2—50-$\mu\mu$fd. smallest fixed mica
C_3—0.25-μfd. tubular, 400 v.
C_4—.05-μfd. tubular
C_5, C_6—25-μfd. 25 or 50 v. electrolytic
C_7—.005-μfd. midget mica
R_1—500,000-ohm ¼- or ½-watt midget resistor
R_2—10,000 ohms, ½ watt
R_3—2000 ohms, ½ watt
R_4—50,000 ohms, 1 watt
R_5—100,000-ohm pot., a.f. (audio gain) taper
R_6—400 ohms, 10 watts
T_1—Small 1-3 interstage a.f. trans.
L—See text

bent back on itself so that the spacing of the two wires is approximately equal to the wire diameter. The grid wire is cut off shorter than the plate wire, in order to allow the insertion of the small grid condenser and grid leak. The overall length of the tank, from the center of the tube caps to the center of the bolt in the standoff insulator which supports the closed end of the "U" and acts as the plate voltage connection is 7⅜ inches. This pillar type standoff insulator is 2 inches high.

Tuning is by means of an improvised split-stator type condenser, the rotor of which is left "floating." A Cardwell ZR-35-AS "Trim Air" is operated upon as follows. Disassemble the condenser so that all rotor and stator plates are removed. Discard all except four rotor and two stator plates. The four remaining rotor

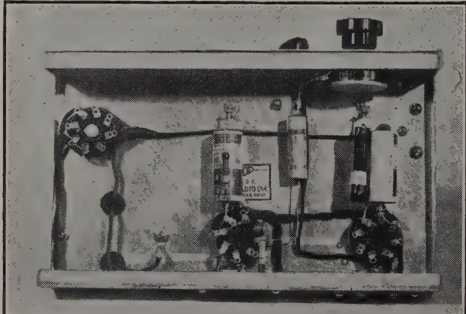


Figure 23.
UNDER-CHASSIS VIEW OF 224-MC. RECEIVER.

plates are not altered, but the two stator plates are trimmed with a pair of heavy shears so that each plate is supported by only *one* of the two stud bolts which originally supported all stator plates. The condenser then is assembled, making use of the original spacing washers, so that the two stator plates are 5⁄16 inch apart, one plate being supported by one stud bolt and the other plate being supported by the other stud bolt. The four rotor plates are then attached, spaced so that each stator plate is enveloped by two rotor plates with the original spacing of .03 inch between adjacent rotor and stator plates. Inspection of Figure 21 shows how the condenser looks when reassembled.

Connection from each stator to the parallel wires is made by means of two ⅞-inch solder lugs, the lugs being bent in towards each other as illustrated in order to permit connection at approximately the same point on each tank wire with respect to the closed end of the tank. The tuning condenser is mounted inverted by means of a "Trim Air" bracket so that the lugs attach to the tank wires 2¾ inches up from the bolt through the bottom of the "U." The condenser is driven by means of an insulated shaft extension.

The antenna coupling loop is made of no. 12 enameled wire, bent as shown in Figures 20 and 21, and varied with respect to the tank wires in order to vary the coupling.

Condenser C_7 should be grounded directly to the chassis with the shortest possible lead.

The receiver runs at full plate voltage at all times, the antenna coupling being adjusted to

the closest value which will still permit superregeneration.

When the receiver is initially put into operation, the frequency range should be checked by means of Lecher wires. If slightly off, the frequency range can be altered sufficiently by varying the spacing between the two tank wires: spreading the wires slightly *lowers* the frequency. If the frequency is very far off, it will be necessary to alter the length of the tank wires slightly as required to enable the tuning condenser to cover the band.

112-MC. MOBILE TRANSCEIVER

With a few minor circuit changes and additional components, the 112-Mc. superregenerative receiver illustrated in Figures 18 and 19 makes an excellent transceiver for mobile work. An output of between 2 and 3 watts, enough to deliver a strong signal over considerable distance, is obtainable at the maximum recommended plate voltage.

The layout is substantially the same as that for the receiver, illustrated in Figure 18, and therefore is not shown here. Also, the remarks pertaining to the r.f. portion of the circuit, including tank coil, tuning condenser, and adjustable antenna coil, apply to the transceiver.

Dual volume controls are provided to permit independent adjustment of gain when receiving and when transmitting. Microphone voltage is obtained from a standard 1½-volt dry cell, in order to avoid the possibility of vibrator or generator hash getting into the speech system through the 6-volt supply lead. The battery also provides C bias for the 6J5 speech or audio amplifier. Because the drain on the battery is so low, many hundreds of hours of transmission are possible before replacement is required.

To prevent a.f. feedback it may be found necessary to neutralize the capacity which exists between contacts on the send-receive switch. Should the a.f. system go into oscillation when the gain control is advanced,

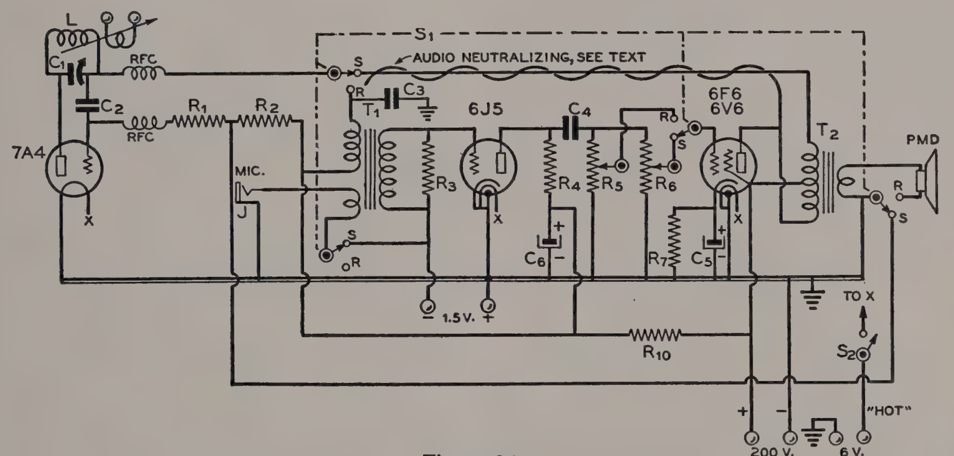


Figure 24.

112-MC. MOBILE TRANSCEIVER SCHEMATIC DIAGRAM.

C_1—5-μμfd. double-spaced midget condenser (with mounting bracket and ceramic shaft coupling)
C_2 — 100 - μμfd. smallest mica condenser
C_3—.01-μfd. tubular condenser, 400 v.
C_4—.05-μfd. tubular condenser, 400 v.
C_5—10-μfd. 25 v. electrolytic
C_6—8-μfd. midget tubular electrolytic, 450 v.
R_1—2500 ohms, 2 watts
R_2—1 meg., 1 watt
R_3—100,000 ohms, ½ watt
R_4—100,000 ohms, 1 watt
R_5, R_6 — 100,000-ohm potentiometer, a.f. gain taper
R_7—500 ohms, 5 watts
R_8—7500 ohms, 1 watt
RFC—Midget u. h. f. chokes
J—Open circuit jack (a closed circuit jack will short the microphone battery)
S_1—4-pole 2-throw rotary switch
S_2—S.p.s.t. toggle switch
T_1 — Transceiver transformer: plate and s.b. mike to single grid
T_2—Universal output transformer: 14,000-ohm c.t. pri., adjustable voice coil winding
PMD—Small p.m. dynamic speaker

simply run a length of insulated wire from the plate of the output tube to the switch, this wire being twisted around the wire running from the opposite end of the transformer to the switch. At the switch, the end of the free wire is adjusted with respect to the wire from T_1 (thus varying the capacity between them) until it is possible to run the gain full on, both on transmit and on receive, without a.f. feedback.

Occasionally such feedback can be eliminated simply by transposing the two secondary wires on T_1, in which case the neutralizing lead will not be required.

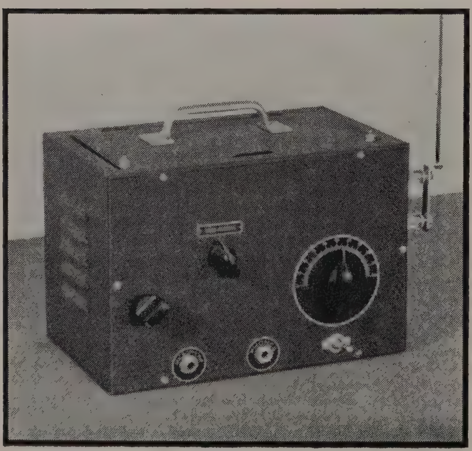
The two r.f. chokes should have their leads clipped off short on the "hot" end to minimize the length of connecting wire between r.f. chokes and the tank circuit.

The adjustable antenna coupling serves as regeneration control, the detector running at high plate voltage at all times. The coupling always is adjusted to the closest value which will still permit superregeneration. This provides maximum sensitivity when receiving and maximum output when transmitting.

The plate supply voltage should not greatly exceed 200 volts on transmission, as excessive plate voltage will cause the 7A4 to overheat and the plate current to "run away."

An antenna system suited for mobile use with this transceiver is described in Chapter 22. The distance which can be worked depends upon the antenna and the location; 50 miles is common from an elevated location.

Figure 25.
114-MC. SELF-CONTAINED TRANSCEIVER.
Two of these battery powered transceivers can communicate over a line-of-sight distance of approximately 20 miles.



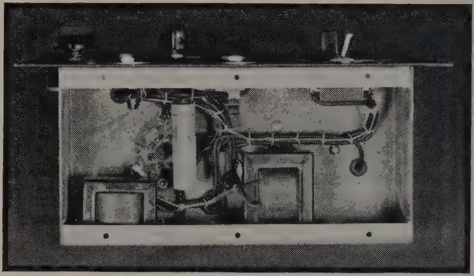


Figure 26.
UNDER-CHASSIS VIEW OF THE TRANSCEIVER.
Layout of parts in the a.f. section is not critical, and that shown here need not be adhered to. The phone jack is insulated from the front panel, the microphone jack is not.

SELF-CONTAINED BATTERY-POWERED TRANSCEIVER FOR 112 MC.

The small transceiver illustrated in Figures 25–29 will provide reliable communication over a distance of 20–25 miles when both stations are sufficiently elevated as to be within line of sight. It is entirely self-contained, being powered by standard portable batteries, and weighs about 11 lbs. (not including microphone or earphones). Cost of operation will be about ½ cent per hour, the battery drain being quite low. The set quite easily can be made even lighter and more compact, but a smaller battery pack will be required. As the smaller packs cost as much in spite of their shorter life, a compromise between portability and economy of operation has to be made.

Construction. The transceiver is constructed in a standard, manufactured cabinet measuring 7 x 10 x 6 inches deep, a sub-chassis measuring 4 x 8 x 2 inches high being supported from the front panel as shown in the illustrations. These illustrations give a good idea of how the parts are arranged to provide an efficient and handy layout. The chassis is sufficiently small to allow the various batteries to be placed between the chassis and the walls of the cabinet. Small pieces of wood or heavy cardboard are wedged between the chassis and batteries to hold the latter firmly in place, a necessary requirement if the unit is to be used while in motion.

As both sides of the tuning condenser are "hot," the shaft is driven through a ceramic coupler. It is preferable to have the grid condenser go to the rotor rather than the stator of the tuning condenser. The ceramic tube socket, the tuning condenser, the grid condenser, and

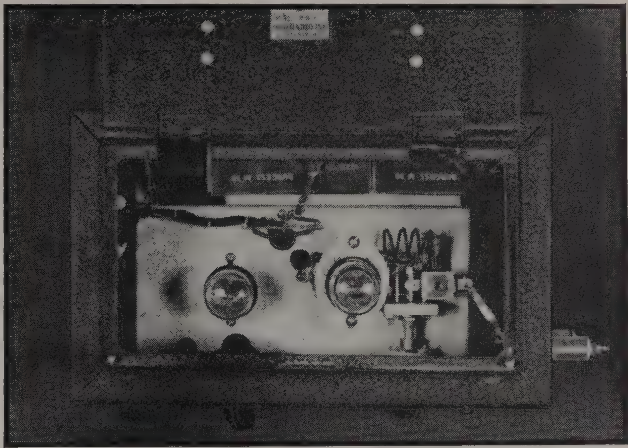


Figure 27.

LOOKING DOWN INTO 114-MC. TRANSCEIVER CABINET.

The various batteries are arranged around the space between the chassis and cabinet, and held firmly in place by wedges of wood and cardboard. A midget 4½-volt C battery, a small 1½-volt A battery, and two portable size 45-volt B batteries make up the battery compliment.

the coil are all placed as close together as possible to provide the shortest possible leads. The coil is soldered directly to the condenser terminals. The "hot" lead wire on each r.f. choke is cut off short before wiring the choke in the circuit.

The antenna consists of a vertical half-wave rod, capacitively coupled to the grid of the

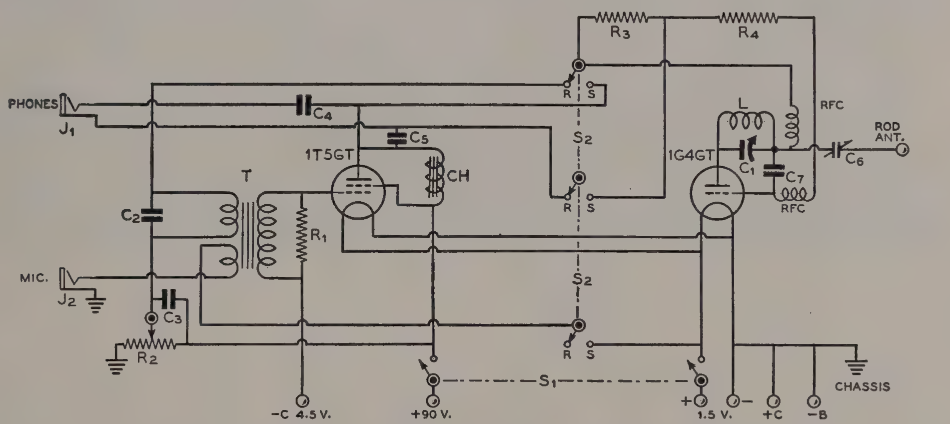


Figure 28.

WIRING DIAGRAM OF THE SELF-CONTAINED TRANSCEIVER.

C_1 — Sub-midget variable condenser with all but two plates removed, about 5 $\mu\mu$fd. Rotor insulated from frame. Connect rotor to grid, stator to plate. Ceramic coupling must be used on shaft

C_2 — .002-μfd. midget mica

C_3—0.1-μfd. tubular, 200 or 400 v.

C_4—0.1-μfd. tubular, 200 or 400 v.

C_5—.01-μfd. tubular, 400 v.

C_6 — 3-30 $\mu\mu$f.d. mica trimmer, ceramic insulation, screw removed. Soldered directly to rotor terminal of C_1

R_1—100,000 ohms, ½ watt

R_2 — 100,000-ohm potentiometer

R_3 — 1 meg., ½ watt

R_4 — 25,000 ohms, ½ watt

CH—Midget 7 to 10 hy. choke, 15 ma. or more

T — Transceiver type midget dual purpose a.f.t., plate and single button mike to grid

S_1—D.p.s.t. toggle switch

S_2—4-pole 2-throw rotary "send-receive" switch

J_1, J_2—Open circuit jacks

RFC—U.h.f. type radio frequency chokes

L—4 turns no. 14 enamelled wound on ½" dia., spaced to hit 114 Mc. with C_1 set at half capacity

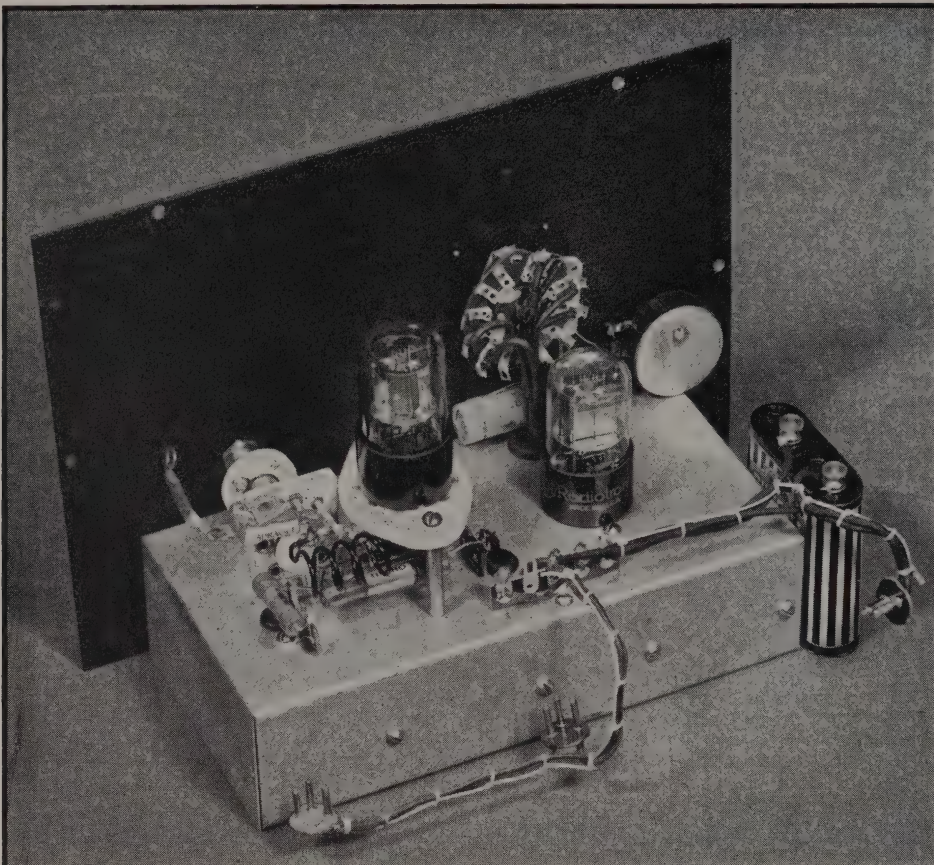


Figure 29.
INTERIOR CONSTRUCTION OF BATTERY TRANSCEIVER.

tube. Better results, both receiving and transmitting, are obtained with the antenna coupled to the grid rather than the plate. The length of the antenna, *overall,* from the tip of the rod to the coupling condenser C_6 should be about 3 feet 6 inches. This is not quite as long as the usual 114-Mc. dipole, but it is an electrical half wavelength just the same because of the loading effect of condenser C_6.

The antenna rod proper is about 3 feet 3 inches long. The rest of the length is made up by the feed-through insulator bolt and the flexible lead to the coupling condenser. Two medium-sized feed-through insulators are mounted one above the other to support the antenna rod, as shown in the illustration. The top insulator does not connect to anything; it merely serves to hold the antenna rod vertical. The two threaded rods for the feed-through insulators should be no longer than necessary, in order to reduce the stray capacity to ground (cabinet) as much as possible.

The antenna coupling is varied by adjusting the distance between the movable and the stationary plate. Closer spacing provides tighter coupling. The coupling should be increased to as much as will still permit superregeneration over the entire band, and then left alone. Ordinarily this adjustment will be about the same as the position assumed by the movable plate when the adjusting screw is removed. When the coupling is increased until a "dead spot" appears, it should occur near the center of the band (114 Mc.). If it occurs at a different frequency, the antenna length should be altered accordingly.

To permit carrying of the transceiver by means of a drawer pull fastened to the lid, the lid is fastened down by means of removable thumb screws. To facilitate carrying the unit when space is limited, the antenna rod is readily detachable.

CHAPTER NINETEEN

U.H.F. Transmitters

The frequencies above 30 megacycles are generally called the ultra-high frequencies or the ultra-short wavelengths. Four amateur bands fall on frequencies above 30 Mc.; the 56 to 60 Mc., 112 to 116 Mc., 224 to 230 Mc., and 400 to 401 Mc. bands. Equipment designed for use in these frequency ranges is generally quite different from the equipment designed for use below 30 Mc. Hence, this chapter will deal with the practical design of transmitters for use within the limits of these bands.

The primary activity on the u.h.f. bands is telephony, although some i.c.w. and occasionally some c.w. is heard. On the 5-meter band (56–60 Mc.) radiophone transmitters are either crystal controlled or m.o.p.a. with a very high-Q self-excited oscillator and preferably at least one buffer stage. Modulated oscillators are not suitable for use on the 5-meter band, as the stability requirement set forth in the FCC regulations automatically rules them out. Frequency modulated transmission is, however, permitted in the range from 58.5 to 60 Mc., and on all frequencies within the bands above 60 Mc.

On $2\frac{1}{2}$ meters (112–116 Mc.), $1\frac{1}{4}$ meters (224–230 Mc.), and $\frac{3}{4}$ meters (400–401 Mc.) the FCC is more lenient, and modulated oscillators are permitted in the interest of simplicity. However, some attempt at stabilizing the oscillator is usually made, and the advantages of m.o.p.a. transmitters are the same as on the low-frequency bands, when greatest simplicity is not needed. Oscillator stabilization is usually accomplished through the use of high-Q circuits, particularly in the grid circuit. High Q is obtained through the use of linear tanks (parallel rods or pipes) or by concentric tanks. The circuit Q is often increased still further in the grid circuit by tapping down on the quarter-wave grid line for the grid connection to the tube.

Portable and mobile operation on frequencies above 112 Mc. can be accomplished with a minimum of equipment through the use of transceivers, or combined transmitter-receivers; these have been described in the previous chapter.

Chapter Subdivisions. In order to classify the types of equipment used on the ultra-high frequencies and the micro waves, this chapter will be subdivided into the following divisions: *Oscillators and M.O.P.A. Transmitters, Crystal Controlled Transmitters and U.H.F. Amplifiers, Frequency Modulation Transmitters,* and *Micro-Wave Transmitters.*

OSCILLATORS AND M.O.P.A. TRANSMITTERS

The majority of the equipment to be shown under this heading will be of the simple oscillator type, since this type of equipment is quite adequate for experimental 112- and 224-Mc. communication. However, when greater frequency stability is desired, it is always advisable to place an amplifier or frequency multiplier between the oscillator and the final amplifier which is to be keyed or modulated. Some of the newer u.h.f. triodes such as the HK-24, 35TG, HY-75, and 1628 can be operated quite efficiently as push-pull triplers and will allow quite satisfactory neutralization in a push-pull amplifier when the conventional cross connected neutralizing circuit is used. Single ended amplifier stages can be neutralized most satisfactorily by the *coil* or *inductive* neutralization circuit shown under *Transmitter Theory*. The "coil" in this case can best be a short section of closely spaced open-wire line to resonate to the operating frequency by the grid-to-plate capacity.

U.H.F. Push-Pull Beam Tubes. Within the last year or so several excellent push-pull u.h.f. beam tubes have made their appearance: 829, 815, etc. These tubes make excellent push-pull r.f. amplifier stages at 56, 112, and 224 Mc., and they have the advantage that, if the input circuit is properly shielded from the output, no neutralization will be required.

112-Mc. Equipment

20-Watt HY-75 Oscillator. This little transmitter was primarily designed to replace the final amplifier stage of a 10-meter mobile transmitter and to be modulated by the speech and modulator system which was originally used with the 10-meter transmitter. It consists of an HY-75 ultra-audion oscillator with conventional coil-and-condenser tank circuit. A concentric pipe or parallel rod oscillator would undoubtedly give greater stability, but with a low-capacity high-transconductance tube the stability has been found sufficiently good with the tank circuit shown. The only precautions that need be taken in the construction of the transmitter is to make sure that all r.f. leads are as short as possible, that all parts are mounted rigidly, and that good u.h.f. insulation be used where it is in contact with high potential r.f. The tuning condenser should be of the ultra midget type, and it should be wired so that the rotor goes to the grid. The exact number of turns for the tank coil will depend somewhat on the physical layout and particular make of components chosen. Some pruning may be required on the coil. It should hit the band when the tuning condenser is about half meshed. Observe that both rotor and stator are hot to ground, both to d.c. and r.f.

The tube socket is not exposed to r.f., and may be of the inexpensive wafer type. It is important that the tube be mounted in a vertical position for good filament life.

Figure 1 shows a back view of the oscillator. It has been mounted upon this small chassis so as to take up as little space as possible when placed alongside the modulator system for the mobile 10-meter transmitter. Normal operation of the oscillator will be with 300 volts at about 80 ma. on the plate. If desired, the power input may be raised to 425 volts at 80 ma. to give about 35 watts input. The circuit diagram of the oscillator is shown in Figure 2.

Inexpensive 8-Watt Oscillator and Modulator. For the amateur who wishes to build an inexpensive low-power station transmitter for 112 Mc., the unit shown in Figure 3 and diagrammed in Figure 4 is ideal. It uses inexpensive tubes throughout, is built upon a breadboard, and has a quite respectable power output capability.

The oscillator is built on a baseboard measuring 5 x 23 x $\frac{3}{4}$ inches. The two rods are each 15$\frac{1}{2}$ inches long, of either copper or aluminum $\frac{3}{8}$-inch o.d. tubing. They are supported on 1$\frac{1}{2}$-inch standoff insulators, placed as shown.

The "stock" supporting the tubes is made from a block of wood measuring 4$\frac{1}{4}$ x 2$\frac{1}{2}$ x $\frac{3}{4}$ inches. The grain should run the long way of the block. Holes are drilled just large enough to take the bases of the tubes, their centers 1$\frac{5}{8}$ inches apart and $\frac{7}{8}$ inch from one of the 4$\frac{1}{4}$-inch edges. Now with a ripsaw, cut the length of the block parallel to the long edges, through the centers of the two socket holes. The tubes will be held firmly, with a vise-like

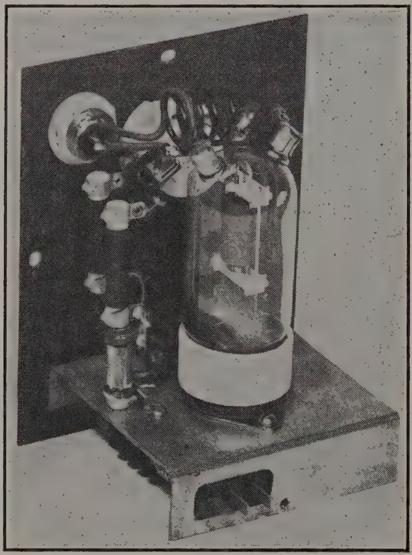


Figure 1.
20-WATT 112-MC. OSCILLATOR.
This diminutive oscillator will take 35 watts input on 112 Mc., and will deliver quite a substantial signal on the band. The tube clips are connected to the tank condenser by means of narrow copper ribbon. A 1-turn link at the grid end of the tank connects to a coaxial cable connector.

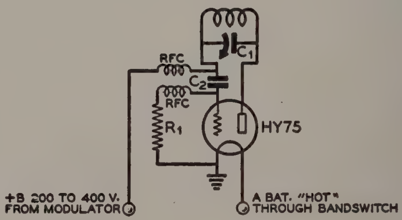


Figure 2.
SCHEMATIC OF THE HY-75 112-MC. OSCILLATOR.

C_1—15-$\mu\mu$fd. sub-midget condenser
C_2—.0001-μfd. midget mica
RFC—U.h.f. choke
R_1—2500 ohms, 1½ watts
Coil—4 t. no. 14 enam., ½" dia. spaced to hit band

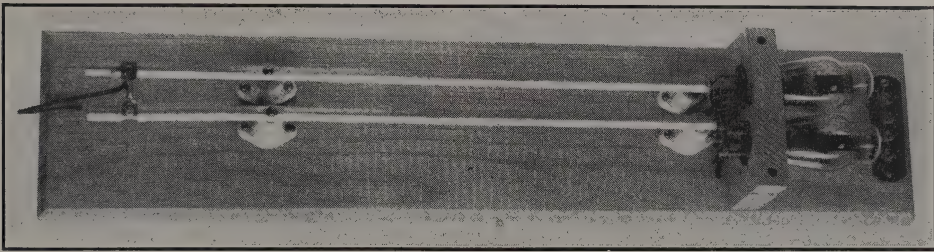


Figure 3.
112-MC. BREADBOARD OSCILLATOR USING PUSH-PULL TRIODES.
The oscillator is built breadboard fashion, with connections made directly to the tube prongs. The linear tank circuit permits comparatively high efficiency and gives an output of 8 to 10 watts. A suitable modulator and power supply are shown in Figure 4. 6P5-GT's or 7A4's may be substituted for the 76's shown if desired; it will be necessary to lengthen the rods about ½-inch with the 6P5-GT's and 7A4's.

grip, when two screws are run down through the assembly and into the baseboard 5 inches from one end of the latter.

This method of mounting the tubes, and soldering direct to the tube prongs, permits shorter leads than could be obtained with any type socket, as even socket terminals represent objectionable lead length at this frequency.

Bakelite tube bases show rather high losses at 2½ meters, but it is possible to reduce these losses by putting two hacksaw slots in the base of each tube, between the plate and grid

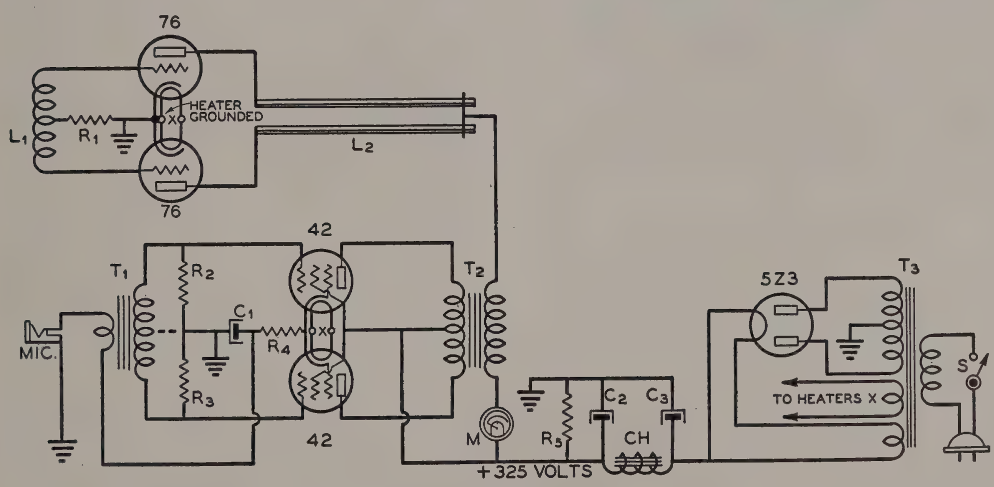


Figure 4.
WIRING DIAGRAM OF THE ECONOMY 112-MC. TRANSMITTER.

L_1—5 turns no. 14 enameled, ½" dia., spaced to approx. ⅜" (see text)
L_2—Copper or aluminum tubing "linear tank," each element ⅜" dia., 15½" long, spaced 1¼" center to center (see text)
C_1—25-μfd. 25-volt electrolytic
C_2, C_3—Single "dual 8-μfd." electrolytic, 450 w.v.
R_1—7500 ohms, 1 watt
R_2, R_3—200,000 ohms, 1 watt
R_4—400 ohms, 10 watts
R_5—50,000 ohms, 2 watts
T_1—High ratio sing. button mike trans. (see text)
T_2—Class B output transformer for 6N7, 6A6, 53, etc. to class C load
T_3—350 v. each side c.t. at 110 ma.; 5 v. at 3 amp.; 6.3 v. at 2 amp.
CH—10 to 30 hy., 110-ma. filter choke
M—0-100-ma. milliammeter
MIC—Closed circuit jack for microphone

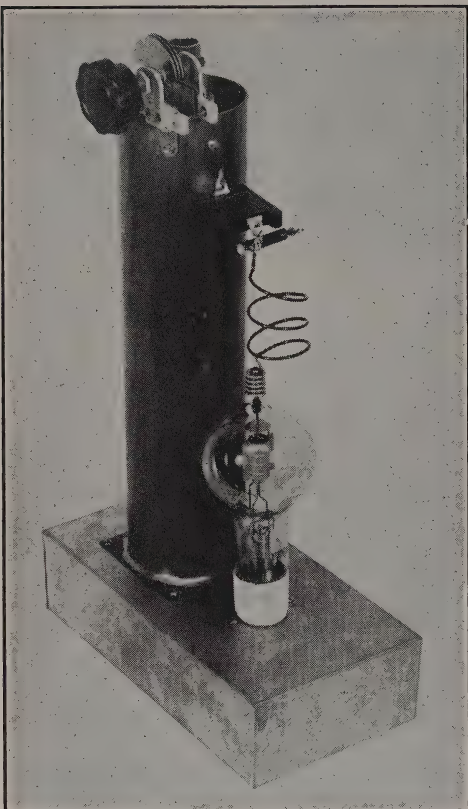
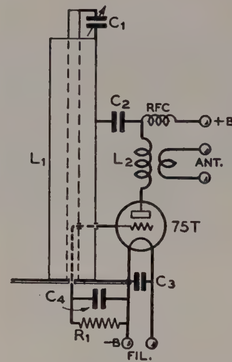


Figure 5.
CONCENTRIC-LINE 75T OSCILLATOR.
This concentric-line oscillator with a 75T gives good stability and a quite reasonable power output on the 112-Mc. band.



C_1—35-$\mu\mu$fd. midget variable
C_2—.0001-μfd. 5000-volt mica
C_3—.003-μfd. midget mica
C_4—.0005-μfd. 1000-volt mica
R_1—5000 ohms, 10 watts
L_1—Concentric grid line—see text
L_2—Resonant plate coil and link—see text

Figure 6.
SEMI-SCHEMATIC OF THE 75T OSCILLATOR.

prongs. Be sure to saw all the way through the base (about $\frac{1}{16}$ inch), but don't go any farther or you may saw into the glass tip that seals the stem of the tube.

The grid coil is soldered directly to the grid prongs of the tubes, which should be mounted with the grid prong (the isolated prong) upward. The coil consists of 5 turns of no. 14 enamelled, spaced to approximately $\frac{5}{8}$ inch. The exact spacing constitutes tuning of the grid circuit. The carbon resistor which serves as a grid leak is mounted vertically between the grid coil and the wood "stock." The top of the resistor is soldered to the center turn of the grid coil (top of the coil) and the other resistor lead is soldered to the jumper which connects the two 76 cathodes.

The sliding jumper for the plate tank is constructed by soldering together two of the older type grid clips which just slip over a $\frac{3}{8}$ inch diameter. These make firm contact to the rods, and can be slid along by pressing upon the two "tongues" while attempting to slide them. The lead from this jumper runs underneath the baseboard *midway between the two tank rods* to prevent unbalancing of the circuit.

Tuning. The oscillator is tuned by placing the shorting bar 14½ inches from the plate end of the plate tank rods. With the antenna disconnected, squeeze the grid coil in and out until the oscillator draws 50 ma. It should be possible to draw small sparks from the plate end of the rods with the tip of a lead pencil, indicating oscillation. The antenna is now coupled to the plate tank by means of a hairpin link, the coupling being adjusted until the oscillator draws 60 ma. Tighter coupling should not be used, as the life of the 76's will be greatly shortened if they are allowed to draw over 60 ma. for any length of time. The output under these conditions will be very close to 8 watts.

The microphone jack, MIC, must be of the closed circuit (shorting) type. Otherwise the low voltage by-pass condenser, C_1, will be blown when the microphone plug is removed.

100-Watt 75T Resonant-Line 112-Mc. Oscillator. Figures 5 and 6 illustrate a concentric-line controlled 112-Mc. oscillator using a 75T, which will put out approximately 100 watts of stabilized r.f. on any frequency in the 112–116 Mc. amateur band. A short concentric line, which is resonated to the operating frequency by means of a 35-$\mu\mu$fd. midget variable, acts as the frequency determining element; output power is taken from a self-resonant coil in the plate circuit.

The concentric line itself is 12 inches long and $2\frac{7}{8}$ inches inside diameter (3 inches o.d. with $\frac{1}{16}$-inch wall), and the inner conductor

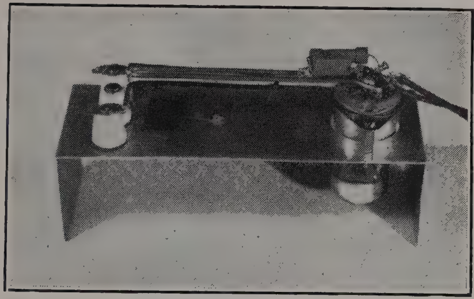


Figure 7.
2-WATT 224-MC. OSCILLATOR.
This simple and inexpensive oscillator may be used either as a low-power transmitter on the 1¼-meter band or, by a slight circuit alteration, as a 224-Mc. band superregenerative receiver.

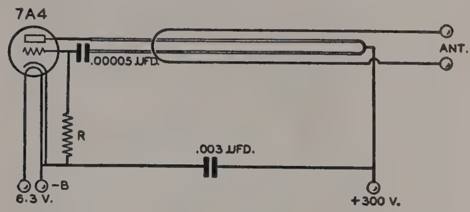


Figure 8.
SCHEMATIC OF THE SIMPLE 7A4 224-MC. OSCILLATOR.
The resistance R should be 7500 ohms for operation of the oscillator as a transmitter. For operation as a superregenerative detector, R should be removed and a 500,000-ohm resistor placed across the grid condenser. The plate circuit of the 7A4 may then be fed into a conventional audio amplifier.

is 13¼ inches long and ¾ inches in diameter. Both pieces which make up the line are cut from standard lengths of thin-wall copper water pipe. To make up the line first the inner conductor is soldered to the center of a piece of 20-gauge copper sheet about 3½ inches square, with the aid of a small alcohol torch and a soldering iron. Then the outer conductor is slipped over it and also soldered in place. Considerable heat is required to do the soldering, but if the work is placed on a block of wood as insulation, a small alcohol torch *and* a conventional electric soldering iron will do the job quite easily. The wood will be thoroughly charred when the work is finished but it will have served its purpose. Asbestos would probably be better but wood will be satisfactory.

A hole is drilled in both the inner and the outer conductor 2¼ inches up from the base on the line. Then another hole is drilled in the center of the base so that a wire may be run through it, through the inner conductor, and then through the hole 2¼ inches up through both the inner and outer conductor to connect to the grid of the tube. This wire is bypassed immediately to ground and one side of the filament of the 75T as it leaves the base of the line.

The plate coil consists of 3 turns of no. 12 wire 1¼ inches in diameter and 2 inches long. The upper end of this coil is by-passed to the concentric line by means of a .0001-µfd. 5000-volt mica condenser. This plate coil was found to resonate over the entire 2½-meter band with the plate-to-ground capacity of the 75T and the distributed capacity of the circuit.

With the circuit constants shown, the grid condenser will tune the oscillator to the center of the 2½-meter band when it is about half meshed. About 30° rotation of the condenser will cover the band. Approximately 100 watts output may be obtained from the oscillator at 1250 plate volts, and at a plate efficiency of 50 to 65 per cent.

224-Mc. Equipment

Within the last few months a great deal of interest has been centered on the 224-Mc. band due to the peculiar conditions which exist upon it, and due to the fact that, for a given amount of power, greater signal strength is obtainable over an optical path than with use of any of the lower frequencies.

A 2-Watt 7A4 Oscillator. Figure 7 shows a 224-Mc. oscillator using a 7A4 which can be used either as a transmitter to give about 2 watts output, or as a superregenerative detector to feed an audio amplifier as a receiver. The unit as shown, and as illustrated in the circuit diagram, is set up as a low-power 224-Mc. oscillator. For this use, the grid leak R should be 7500 ohms and should be connected between the grid of the 7A4 and ground. For the proper method of tuning this oscillator to a given frequency in the 1¼-meter band through the use of Lecher wires, see the chapter *U.H.F. Communication.*

As an oscillator the plate voltage on the 7A4 should be limited to 250 volts and the plate current should not be greater than 30 ma. The resting plate current of the oscillator, unloaded, will be about 18 to 20 ma.; when the circuit is loaded to 30 ma. about 2 watts may be taken from the antenna coupling link.

The plate hairpin of the oscillator is made from no. 10 bare copper wire (actually no. 10

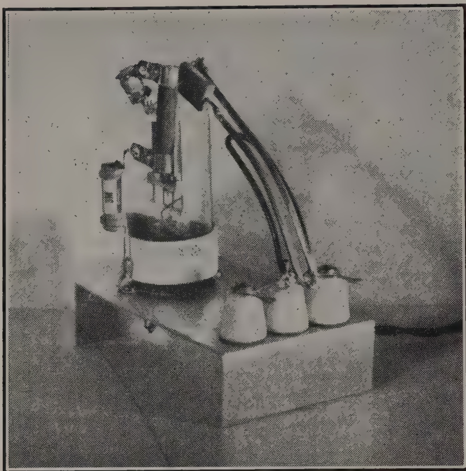


Figure 9.
8-WATT HY-75 224-MC. OSCILLATOR.
This small HY-75 transmitter is ideal for the amateur who wishes a medium-power 1¼-meter oscillator for both fixed station and portable use.

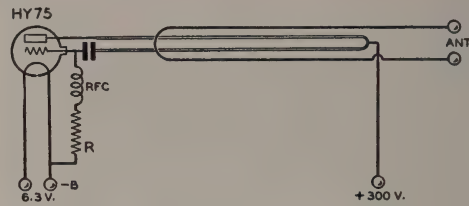


Figure 10.
SCHEMATIC OF THE HY-75 224-MC. OSCILLATOR.
The grid-leak R should have a resistance of about 3000 ohms for normal use. The grid condenser should have a value of .00005 μfd.

enamelled wire from which the enamel has been scraped); it is bent into a narrow hairpin with about $\frac{3}{32}$-inch spacing between the wires. The length from the turn on the loop where the plate voltage connection is made to the plate of the tube is $4\frac{1}{2}$ inches. The length along the other side of the loop from the plate voltage connection to the grid *condenser* is $3\frac{1}{4}$ inches. Quite a wide adjustment in frequency may be obtained by varying the spacing between the wires in the hairpin. Decreasing the spacing *increases* the frequency, and increasing the spacing decreases the frequency of oscillation. It is quite simple to vary the frequency of oscillation from about 180 Mc. up to 230 Mc. merely by making a comparatively small adjustment in the spacing from just over $\frac{1}{8}$ inch to $\frac{3}{32}$ inch.

To convert the oscillator into a superregenerative detector, it is only necessary to remove the 7500-ohm resistor that goes from the grid to ground and then to place a 500,000-ohm resistor directly across the grid condenser. Making the return of the grid leak to positive high voltage in this manner greatly increases the output of the tube when operating as a detector, as compared to when it is returned to ground. Note that it is necessary to have the .003-μfd. by-pass condenser from the plate return to ground for the tube to superregenerate.

An HY-75 8-Watt Oscillator. Another 224-Mc. oscillator using a hairpin as the resonant line is illustrated in Figure 9 and diagrammed in Figure 10. The lead lengths from the center of the hairpin to the plate of the HY-75 and to the grid condenser are the same as for the 7A4 oscillator just described. An r.f. choke has been used between the grid and the grid-leak because of the comparatively low value of resistance of this leak resistor. It was not required in the 7A4 oscillator because of the considerably higher grid-leak resistance. A grid-leak resistance from 3000 to 4000 ohms has been found to be best for the HY-75 in this circuit.

The operating voltage on the HY-75 should be from 275 to 300 volts. The unloaded plate current of the oscillator will be about 30 to 35 ma., and it can safely be loaded to 75 or 80 ma. before excessive plate heating takes place. With this value of power input, the output will be from 8 to 10 watts.

A Push-Pull HY-75 Oscillator. The unusual parallel-rod push-pull oscillator shown in Figure 11 and diagrammed in Figure 12 has proven to be quite a satisfactory source of power for experiments in the 224- to 230-Mc. amateur band. A parallel-rod line is used as the frequency controlling element, and a small self-tuned coil is used in the plate circuit. The resonant line is made up of two $\frac{7}{8}$-inch, thin-wall copper pipes spaced $\frac{7}{8}$-inch, $9\frac{1}{2}$ inches long overall, and connected together both at the top and bottom to act as a half-wave line instead of the more common quarter-wave arrangement. The base for the line is a piece of 20-gauge sheet copper $1\frac{3}{4} \times 4$ inches which is mounted above the $9\frac{1}{2} \times 5 \times 1\frac{1}{2}$-inch chassis by means of $\frac{1}{2}$-inch stand-off insulators.

The capacity to chassis of the copper base plate acts as a by-pass for the center of the parallel-rod line. The copper plate can be proved to be acting normally as a by-pass,

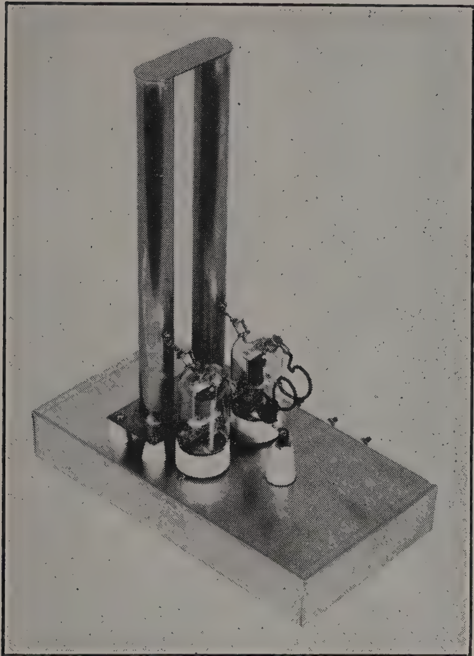


Figure 11.
PUSH-PULL 224-MC. HY-75 OSCILLATOR.

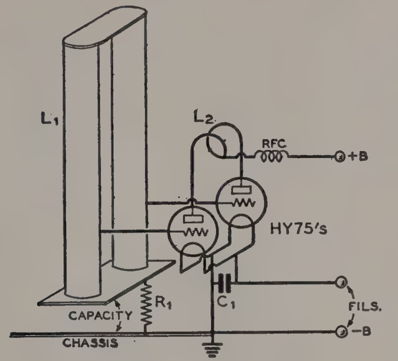


Figure 12.
SEMI-SCHEMATIC OF THE PUSH-PULL 224-MC. OSCILLATOR.

C_1—.003-μfd. midget mica
R_1—5000 ohms, 10 watts
L_1—Half-wave parallel-rod line
L_2—2 turns ⅝" dia., 1" long
RFC — 6 turns hook-up wire, ¼" dia.

since its center will be quite cold to r.f. One of the standoffs which supports the copper plate is of the feedthrough type, and has the grid-leak connected between its lower end and the grounded side of the filaments of the tubes.

The power output of the oscillator as shown is 20 to 25 watts, with 450 volts on the plates of the tubes. The plate efficiency is approximately 40 per cent with the half-wave line in the grid circuit as shown. The plate efficiency was somewhat less than this until the original quarter-wave grid line was replaced with the capacity-shortened (grid-to-ground capacity) half-wave line.

50-Watt 225-Mc. 829 M.O.P.A. Transmitter. Figures 13 and 14 illustrate a very interesting 225-Mc. transmitter of quite respectable power handling capabilities. This transmitter is particularly interesting in the fact that it is an oscillator-amplifier affair instead of being merely an oscillator, as are most transmitters for this high a frequency. The fact that the final stage is an amplifier indicates that it is quite possible to double down to a frequency as high as 225 Mc. for crystal controlled or frequency modulation transmission and still be able to find an arrangement which will be capable of operating as an amplifier at this extremely high frequency. As a matter of fact, the 829 amplifier stage operates with a plate efficiency of about 60 per cent when fully loaded, and requires a driving power of less than 5 watts actual output from the preceding stage.

The 829 tube itself is particularly designed for operation as an r.f. amplifier for frequencies above 50 Mc. It consists of a pair of beam tetrodes with a total plate dissipation of 40 watts mounted inside an envelope in which lead length has been made a primary consideration. The tube has no base, the terminal leads for the tube elements being brought out to tungsten rods which extend through the glass bottom plate of the envelope.

The socket for this tube is also very interesting and it, in addition, is particularly designed for u.h.f. use. The photographs give a good general idea of its construction: all the leads which are normally cold, heaters, cathode, and screens, are brought out through large terminal clips which have built-in mica by-pass capacitors. Then, the grid leads to the two elements within the envelope are brought out to a separate Mycalex arbor which is supported away from the base of the socket by means of small ceramic pillars.

The general layout of the HY-75 oscillator which is used as the exciter for the 829 can be seen in the top view photograph. The oscillator circuit is an ultra-audion, with a combination resonant line and coil in the plate circuit. The lead from the plate extends about 1½ inches and the lead from the grid condenser

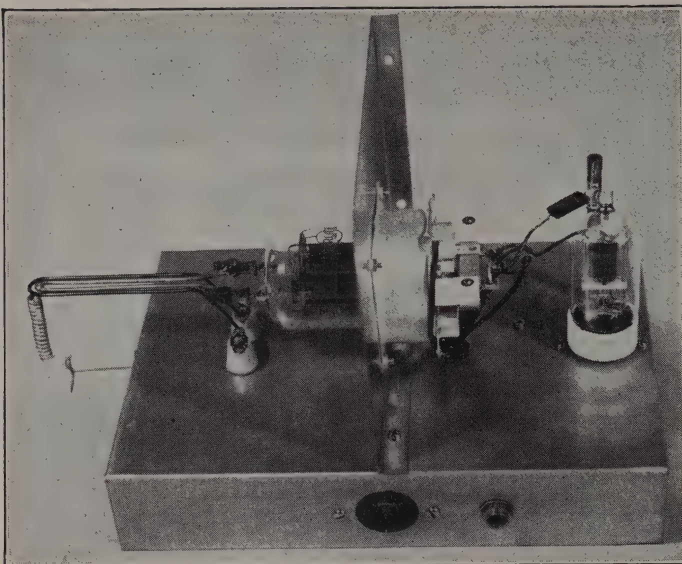


Figure 13.
TOP VIEW OF THE 829 M.O.P.A. 225-MC. TRANSMITTER.

about ½ inch and then they are crossed over to form a 1-turn coil. Another 1-turn coil is interwound with this and connected to the two grid terminals on the 829 socket. The schematic diagram, Figure 15, gives a general idea of the arrangement of these two circuits but it does not indicate graphically the fact that the two 1-turn coils are interwound—at least in as much of a manner as two 1-turn coils can be interwound.

If desired, the frequency of the HY-75 oscillator may be controlled by a quarter-wave concentric line, in the same general fashion as the frequency of the 75T 112-Mc. oscillator is controlled. The grid of the HY-75 should be tapped up a short distance from the bottom of the capacity loaded line, and the plate return made to the side of the line in the same manner as the 75T oscillator described previously. An alternative arrangement would be to use the HY-75 as a frequency doubler from the 112-Mc. band for crystal controlled or f.m. transmission. The plate and grid circuits of the 829 amplifier would be the same as shown, and the plate tank of the HY-75 would be returned to ground with the 112-Mc. excitation fed to the grid.

The normal plate voltage of the 829 is 400 volts, the screen voltage is 200 volts, and the grid bias should be 35 to 45 volts. The grid current of the 829 as shown is about 8 to 9 ma. through a 4000-ohm grid-leak. The amplifier operates very satisfactorily with a plate current of 200 ma., and the plate current may be run as high as 240 ma. if the full rating of the tube is to be used. The output at 200 ma. plate current (400 plate volts) is about 40 watts; at 240 ma. plate current it is about 50 watts.

The total length of the no. 10 bare wire tank circuit is 4¼ inches from the plate seals of the tube to the end of the hairpin. The spacing between wires is ⅛ inch for about 3 inches until the wires spread out to make soldered connection to the plate clips of the 829. The actual plate clips are small hard copper spring clips of the type supplied with HK24 tubes to make the plate connection to them. The plate line is resonated to the frequency of the oscillator by sliding the line back and forth on the tungsten rods that come out of the 829 envelope as the plate connections. The type of plate clips shown are par-

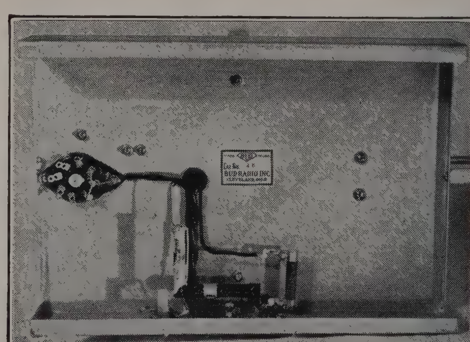


Figure 14.
BOTTOM VIEW OF THE 829 TRANSMITTER.

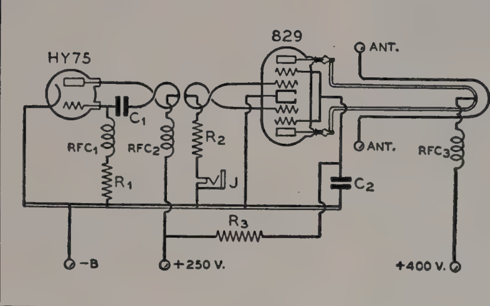


Figure 15.
SCHEMATIC OF THE HY-75, 829 224-MC. TRANSMITTER.

C_1—.00005-μfd. midget mica
C_2—.0005-μfd. midget mica
R_1—7500 ohms, 1 watt
R_2—4000 ohms, 1 watt
R_3—7500 ohms, 10 watts
RFC_1—Midget u.h.f. choke
RFC_2—12 t. hookup wire, ¼" dia.
RFC_3—15 t. no. 14, ⅛" dia.
Coils and lines—See text

ticularly suited to this application, since they slide back and forth comparatively freely on the plate lead rods.

CRYSTAL-CONTROLLED U.H.F. TRANSMITTERS

Crystal control provides the same advantages of excellent frequency stability and reliability on the u.h.f. bands that it does on the lower frequencies. However, due to the relatively greater difficulty of getting amplifier and frequency multiplier stages into operation on the higher frequency bands, crystal control is not widely used except in the case of more elaborate transmitters. High-frequency crystals have made their appearance on the market, but due to their inherent instability, high temperature coefficient, and lack of ruggedness, they have fallen into disuse, and, in fact, have been discontinued by some manufacturers. Hence, for most amateur work, the highest practical operating frequency for the crystal is 7300 kc. From this comparatively low frequency a rather large number of doublers are required to get down to the u.h.f. bands. However, through the use of beam tetrodes, it is possible to obtain comparatively good operation from triplers and quadruplers, thus simplifying the frequency multiplication problem.

20-Watt Crystal Controlled 14-, 28-, 56-Mc. Transmitter or Exciter. The crystal controlled 56-Mc. r.f. unit illustrated in Figures 16 and 17 and diagrammed in Figure 18 uses conventional circuits and low cost parts. With but three stages and a 7-Mc. crystal, it supplies 20 husky watts of crystal controlled, 56-Mc. or 28-Mc. r.f. For 'phone operation the output stage may be modulated by a 25-watt modulator. As an exciter it has sufficient output to drive a 56-Mc. final stage to 200 watts input.

The chassis measures 12 x 7 x 2 inches. As can be seen from the photographs, the tubes are evenly spaced along the center of the chassis. Each plate coil is directly in front of the tube with which it operates. The tank condensers are mounted on the front lip of

Figure 16.
THREE-STAGE 20-WATT CRYSTAL-CONTROLLED 56-MC. EXCITER UNIT.
A 6L6 oscillator on 7 Mc. drives a 6L6G quadrupler, which in turn drives a T21 doubler to 56 Mc. Power output may be taken from any of the three stages by means of a coupling link.

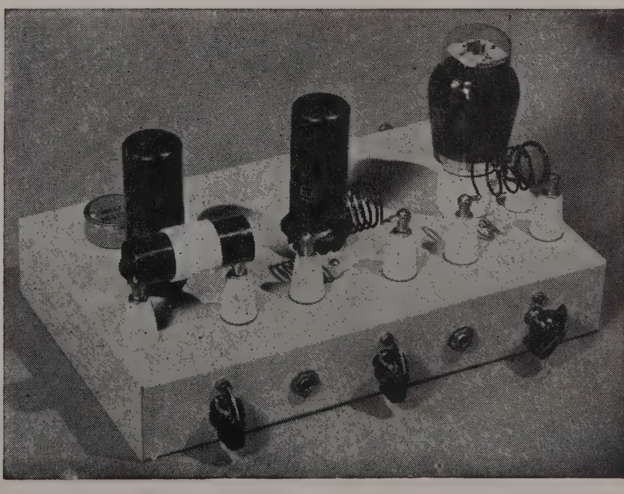


Figure 17.
UNDER-CHASSIS VIEW OF THE 56-MC. R.F. UNIT.

the chassis directly below their respective coils. Small jack type feed-through insulators are used to support the plug-in coils and at the same time to provide connections to the condensers. Banana plugs on the coils allow quick and easy band change.

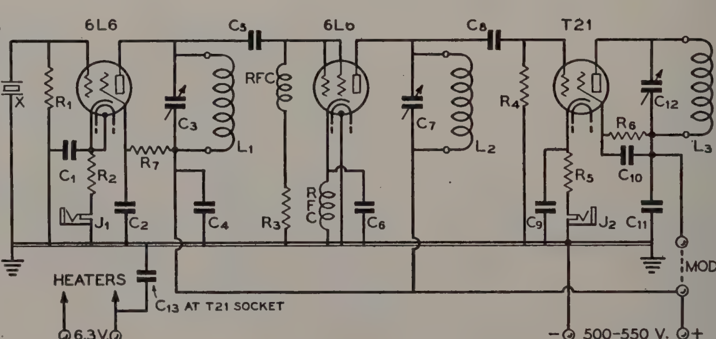
Inasmuch as each tuned circuit is on a different frequency, placing the coils in line along the front of the chassis does not have any adverse effect on the operation of the unit.

Underneath the chassis, parts are placed where convenience dictates. The T21 stage has all its ground return connections made to the feed-through insulator which is at the cold end of the plate tank. While this does not enhance the appearance of the unit, it aids in eliminating coupling in the various ground return circuits.

Two feed-through insulators at the rear of the chassis are provided for the connections from the modulator. If the unit is used as an exciter or c.w. transmitter, these terminals are simply shorted together.

The second 6L6 acts either as a doubler or quadrupler, depending upon the crystal frequency and desired T21 output frequency. Thus, with a 40-meter crystal, 10- or 5-meter output is obtainable from the T21. With an 80-meter crystal, either 20- or 10-meter output is obtainable from the T21.

Figure 18.
SCHEMATIC OF THE 14-28-56 MC. R.F. UNIT.



C_1, C_2—.01-μfd. paper
C_3—75-$\mu\mu$fd. midget
C_4—.005-μfd. mica
C_5—100-$\mu\mu$fd. mica
C_6—50-$\mu\mu$fd. mica
C_7—50-$\mu\mu$fd. midget
C_8—.001-μfd. mica
C_9—.002-μfd. mica
C_{10}, C_{11}—.001-μfd. mica
C_{12}—15-$\mu\mu$fd. midget, double spaced
C_{13}—.001-μfd. mica
R_1—25,000 ohms, ½ watt
R_2—400 ohms, 10 watts
R_3—25,000 ohms, 10 watts
R_4—150,000 ohms, 2 watts
R_5—600 ohms, 10 watts
R_6—30,000 ohms, 10 watts
R_7—100,000 ohms, 1 watt
RFC—2.5-mh. r.f. chokes
J_1, J_2—Closed-circuit jack
Coils—see coil table

COIL TABLE

All coils have small, banana type plugs spaced 2½ in. 80 and 40 m. coils are wound on bakelite tubing; 20, 10, and 5 m. coils are self-supporting.

80 OSC.
37 turns no. 22 d.c.c. on 1 inch form.

40 OSC. OR DOUBLER
22 turns no. 22 d.c.c. on 1 inch form.

20 DOUBLER
13 turns no. 14 enam. 1 in. dia. spaced to 1½ in.

20 FINAL
17 turns no. 14 enam. 1¼ in. dia. spaced to 1½ in.

10 QUADRUPLER
6 turns no. 14 enam. 1 in. dia. spaced to 1 in.

10 FINAL
8 turns no. 14 enam. 1 in. dia. spaced to 1¼ in.

5 FINAL
4 turns no. 14 enam. ⅞ in. dia. spaced to 1¼ in.

Note: 40 meter coil serves either as osc. coil or doubler coil.

With the meter plugged in the cathode circuit of the T21, the total plate, screen, and grid current is shown. This gives a false indication as to the plate current "dip" of the stage, which is about 15 milliamperes lower than the cathode current would indicate.

For optimum performance, the T21 stage should be loaded to approximately 90 milliamperes. At this input, the output is approximately 20 watts.

No antenna coupling circuit has been provided as the type of coupling circuit will depend upon the antenna used. Any of the usual capacitive, inductive, or link-coupling circuits will be suitable, however. When used as an exciter, the unit should be link-coupled to the next stage.

Medium Power 56-Mc. Amplifier. By using tubes having close element spacing, yet low interelectrode capacities, and a plate tank condenser especially designed for u.h.f. service, it is possible to construct a medium power 56-Mc. amplifier that will exhibit good efficiency without resorting to the use of parallel rods in the plate circuit.

Such an amplifier is illustrated in Figure 20. It utilizes a pair of HK-24's in push-pull, and the efficiency is as good as that obtained with commonly used equipment on the 14-Mc. band. With proper coils, the amplifier could also be used on 28 and 14 Mc., but as it was expressly designed for 56-Mc. work, the coils are not of the plug-in type. By fastening the plate coil directly to the condenser stator lugs, losses are minimized.

About 20 watts excitation are required, this amount of excitation permitting approximately 175 watts input on 'phone or 225 watts input on c.w. The T21 exciter of Figure 18 is ideally suited for use with this amplifier, the excitation being sufficient so long as the coupling link between exciter plate coil and amplifier grid coil is not too long. The losses are high at 56 Mc. in a twisted pair line, even in a good line. EO-1 cable makes the best coupling line, and it should be not more than 18 inches long unless reserve excitation is available to compensate for the losses in the line.

A conventional resistor-biased circuit is used with circuit balance provided by a grounded-rotor grid condenser. Plate voltage is fed to the center of the plate coil through a u.h.f. choke. Since the circuit is balanced by grounding the rotor of the grid condenser, it is possible to let the rotor of the plate condenser "float," thus increasing the allowable plate voltage for a given condenser spacing. No filament by-pass condensers are used, as they were found to be unnecessary. Mechanically, the amplifier differs somewhat from the usual push-pull stage, and the mechanical layout will therefore be discussed in greater detail.

Construction Details. An 11 x 7 x 2-inch chassis allows ample room for all the components except the filament transformer, which is mounted externally.

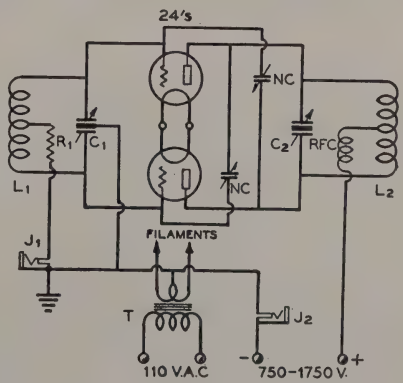


Figure 19.
125-WATT HK-24 U.H.F. AMPLIFIER.

C1—30-μμfd. per section midget
C2—35-μμfd. per section, 4500-volt spacing
R1—3000 ohms, 10 watts
J1, J2—Single closed circuit jacks.
T—Filament transformer, 6.3 v., 6 a.
NC—See text
L1, L2—See text
RFC—U.h.f. choke

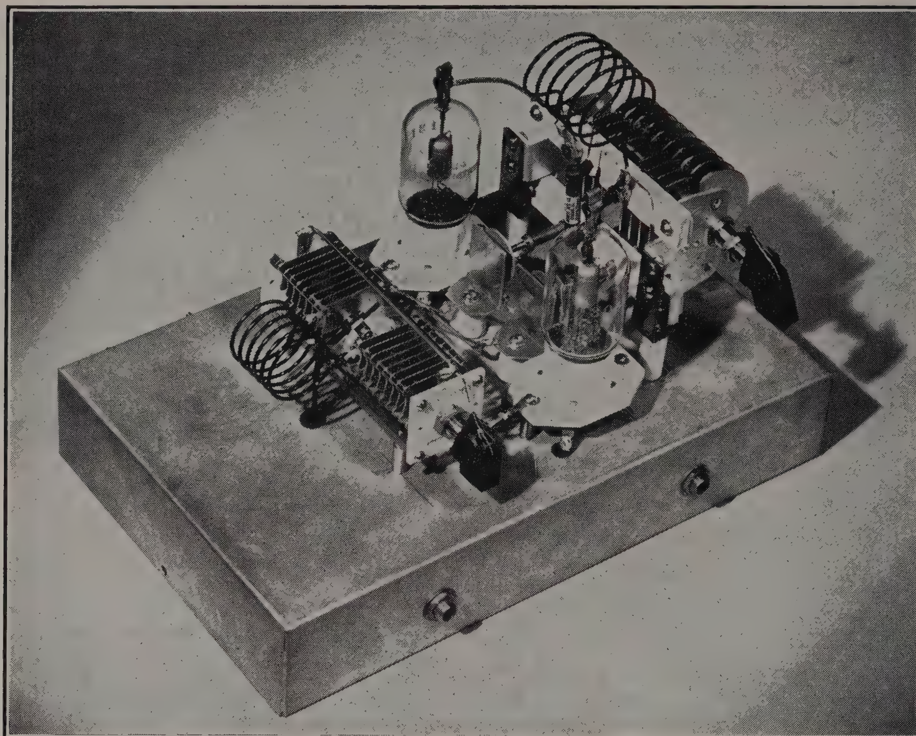


Figure 20.
125-WATT 56-MC. AMPLIFIER.
Extreme simplicity characterizes this 56-Mc. amplifier stage. The neutralizing condensers may be seen between the tubes. All components with the exception of the grid resistor are above the chassis.

The plate condenser is one designed for u.h.f. use. The stator terminals are arranged so as to allow an extremely compact neutralizing condenser assembly. This condenser is mounted on its side with the stator terminals toward the tubes. Two angle brackets and small stand-off insulators serve to hold the condenser above the chassis. Mounting the condenser in this manner permits short plate leads to the upper stator terminals. The plate coil, 6 turns of no. 14 wire $1\frac{1}{4}$ inches in diameter, is spaced so as to mount directly on these upper terminals.

Two small discs of aluminum, 1 inch in diameter and 1/16-inch thick, are used for the movable plates of the neutralizing condensers. Each of these plates has a flat-headed 6-32 screw through its center. The screws are held in place by nuts on the back of the discs. The heads are filed smooth with the surface of the discs. The edges of the discs are rounded with a fine-tooth file to prevent corona losses.

Two pieces of hollow rod, threaded with a 6-32 tap, are mounted on the lower stator terminals of the plate condenser. The screws through the discs are screwed into these rods and neutralizing adjustments are made by running the screws in or out of the threaded rods, thus changing the spacing between the circular plates and the stationary plates, which are simply small rectangular pieces of aluminum mounted on stand-off insulators.

The grid coil is 6 turns of no. 14 enamelled wire $1\frac{1}{8}$ inches in diameter and $1\frac{1}{8}$ inches long. This condenser tunes with its plates about one-third meshed. Both ends of the rotor are grounded for the sake of symmetry.

The amplifier should not be operated for any length of time with the load removed, as the heavy r.f. field within the plate coil will heat and melt the soldered connection at its center. With the tank circuit loaded, however, no trouble of this kind will be experienced.

By slightly exceeding the plate voltage rating and operating the two tubes at 1750 volts, an output of slightly over 200 watts is ob-

tained from the amplifier at the normal plate current of 150 ma. for the two tubes. For modulated operation, the plate voltage should be lowered to 1250 volts, however. Two jacks, J_1 and J_2, are provided for reading the grid and plate current. A 1-turn link is used between the amplifier and the exciter, and the grid current is adjusted to 50 milliamperes under load by varying the coupling.

FREQUENCY MODULATION TRANSMITTERS AND EXCITERS

Frequency modulation, or f.m., transmission is destined to be one of the major uses of the ultra-high frequency bands. The u.h.f. bands are wide enough so that the wide band of frequencies required for f.m. are amply contained. In addition, a practically infinitesimal amount of modulating power is required to modulate an f.m. transmitter, regardless of its power output, and, a last advantage, frequency multipliers, or class C or class B amplifiers may carry f.m. r.f., since the amplitude of an f.m. signal is constant. However, a complete explanation of the theory and practice of f.m. has been given in Chapter 9, so this section will be devoted entirely to the description of equipment designed for f.m. transmission.

Frequency-modulated amateur transmission is permitted from 29,250 to 30,000 kc., 58,500 to 60,000 kc., and in all of the 2½-, 1¼-, and the ¾-meter band. Three pieces of equipment are to be described in this chapter: (1) a stabilized reactance-tube frequency-modulated exciter and (2) a crystal-controlled phase-modulated exciter, either of which is suitable as an exciter for a transmitter on any of the amateur bands wherein f.m. is permitted, and (3) a complete 112-Mc. f.m. transmitter with one of the new RCA-815 50-watt push-pull beam tubes in the output amplifier.

The frequency multiplying and amplifying stages which will be required between the two exciters which are described, and the output amplifier, can be perfectly conventional in every respect. However, it is important that each stage carrying frequency-modulated r.f. be tuned carefully to resonance if undesirable amplitude modulation is to be kept out of the transmitted signal.

Stabilized F.M. Exciter

By comparing the frequency of a self-controlled oscillator with that of a crystal oscillator, and holding the difference between the two at a constant value by means of a discriminator-reactance tube combination, it is possible to hold the self-controlled oscillator frequency quite constant under changes in load and voltage. A frequency-modulated exciter employing this principle is illustrated in Figures 21 and 22.

The Circuit. The exciter is intended to replace a 40-meter crystal stage or v.f.o. to excite transmitters operating in the 10- or 5-meter f.m. bands. The oscillator tube is a 6F6 in a conventional electron-coupled circuit. This stage has its grid circuit in the 80-meter band and its plate circuit tuned to the second harmonic, between 7316 and 7500 kc. A con-

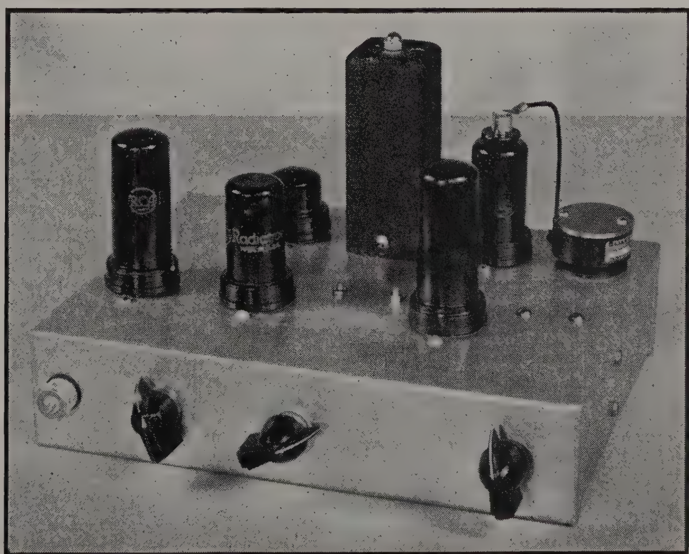


Figure 21.
STABILIZED F.M. EXCITER.
This exciter uses a reactance-tube modulated oscillator with the frequency of the oscillator held at a constant difference from a crystal oscillator. The speech, modulator, and f.m. oscillator section occupies the front portion of the chassis, while the stabilizing circuit is at the rear. The front drop of the chassis carries, from left to right, the microphone connector, gain control, grid condenser control, and plate condenser control.

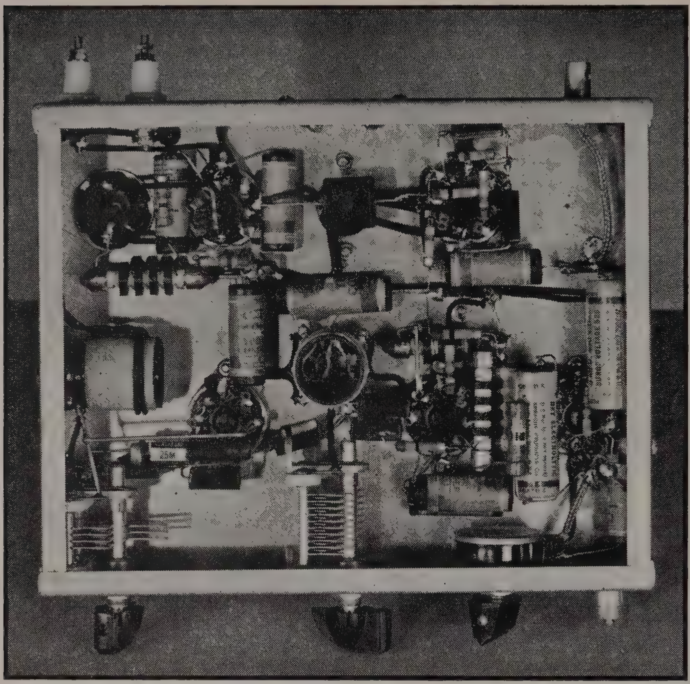


Figure 22.

BOTTOM VIEW OF STABILIZED F.M. EXCITER.

In this photo the oscillator grid and plate coils and condensers may be seen at the center and left edge of the chassis. The speech amplifier is at the lower right, with the reactance tube section to its left. The stabilizing section is at the top, with the converter at the left and the discriminator toward the right.

ventional reactance-tube frequency modulator (operation described in Chapter 9) is tied across the oscillator grid tank.

Instead of going directly through a resistor to ground, the control-grid return of the 6SJ7 reactance tube goes through the two discriminator load resistors, R_{12} and R_{13}, to ground. This allows the d.c. voltage developed by the discriminator to be applied to the reactance-tube grid along with the audio from the speech amplifier. Resistor R_{11} and condenser C_{16} form an R-C filter to remove the audio output of the discriminator.

The 6K8 converter tube receives signal-grid excitation directly from the 6F6 output link, and has a Pierce crystal oscillator circuit in its oscillator section. A signal with a frequency equal to the difference between the crystal and 6F6 output frequencies appears at the 6K8 plate and is applied by means of a discriminator transformer, IFT, to the 6H6 discriminator tube. A voltage which depends upon the difference in frequency between the signal applied to the discriminator transformer and the frequency to which the transformer is tuned, is developed across the discriminator load resistors, R_{12} and R_{13}. When the signal is at the resonant frequency of the transformer, the voltage produced is zero; when the frequency varies one way a positive voltage is produced; a variation in the other direction will give a negative voltage. This voltage is applied as additional positive or negative bias to the reactance tube, and the effect is to cause the reactance tube to restore the frequency back near a value which gives zero voltage output from the discriminator.

Construction. The complete exciter, including the speech amplifier, is built on a 7 x 9 x 2-inch chassis. In Figure 21, the 6N7 2-stage speech amplifier is seen at the left of the front row of tubes, followed to its right by the 6SJ7 reactance-tube modulator and the 6F6 oscillator. The 6K8 and the crystal are at the right rear of the chassis, with the discriminator transformer and the 6H6 to their left. Figure 22 shows how the parts are located under the chassis. The speech amplifier and reactance-tube section is in the lower right in this view, with the oscillator grid tank circuit at the lower center and the output tuned circuit at the lower left. The components at the top of the photo are those associated with the converter and discriminator circuits, with the crystal socket at the left, followed to the right by the 6K8 socket and the 6H6 socket. Two feed-through insulators in the rear drop of the chassis are provided for link connections to the transmitter, while an auto-type connector is used for connection to an external monitoring amplifier. A 4-prong socket is used for connections to the exciter.

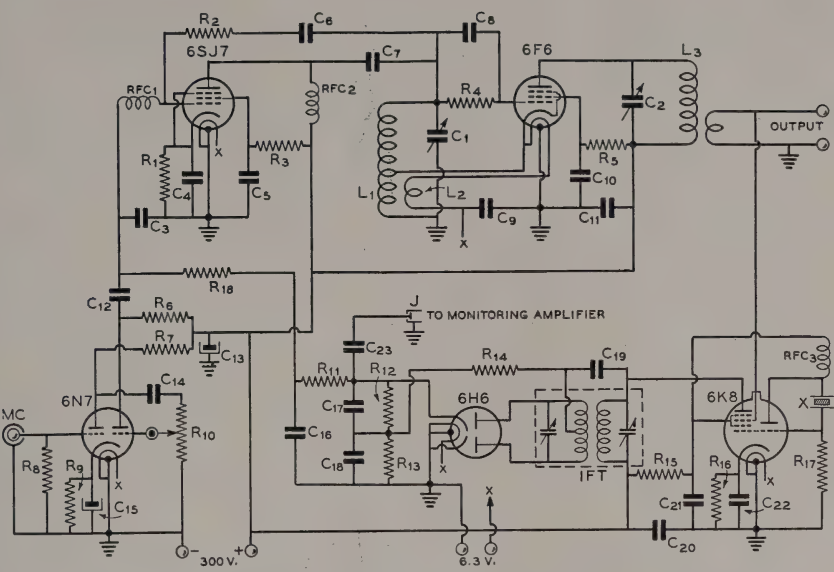


Figure 23.

STABILIZED F.M. EXCITER DIAGRAM.

C_1—140-$\mu\mu$fd. midget variable
C_2—50-$\mu\mu$fd. midget variable
C_3—.0001-μfd. mica
C_4—.01-μfd. 600-volt tubular
C_5—.25-μfd. 600-volt tubular
C_6, C_7—.005-μfd. mica
C_8—.0001-μfd. mica
C_9, C_{10}, C_{11}—.005-μfd. mica
C_{12}—.01-μfd. 600-volt tubular
C_{13}—8-μfd. 450-volt electrolytic
C_{14}—.01-μfd. 600-volt tubular
C_{15}—25-μfd. 25-volt electrolytic
C_{16}—.05-μfd. 600-volt tubular
C_{17}, C_{18}—.0001-μfd. mica
C_{19}—.0005-μfd. mica
C_{20}, C_{21}, C_{22}, C_{23}—.05-μfd. 600-volt tubular
R_1—500 ohms, ½ watt
R_2, R_3, R_4—50,000 ohms, ½ watt
R_5—25,000 ohms, 2 watts
R_6, R_7—250,000 ohms, ½ watt
R_8—1 megohm, ½ watt
R_9—1500 ohms, ½ watt
R_{10}—500,000-ohm potentiometer
R_{11}—500,000 ohms, ½ watt
R_{12}, R_{13}—100,000 ohms, ½ watt
R_{14}—50,000 ohms, ½ watt
R_{15}—30,000 ohms, ½ watt
R_{16}—300 ohms, ½ watt
R_{17}—50,000 ohms, ½ watt
R_{18}—500,000 ohms, ½ watt
RFC_1, RFC_2, RFC_3—2½ mhy.
IFT—465-kc. "output" i.f. trans. with center-tapped secondary
L_1—27 turns of no. 22 d.c.c. close-wound on 1" dia. form, tapped at ninth turn
L_2—9 turns of no. 22 d.c.c. over bottom end of L_1
L_3—25 turns of no. 22 d.c.c. close-wound on 1" dia. form. 2-turn link over "cold" end.
X—Crystal near 7000 kc. for 5- and 10-meter operation, near 7500 kc. for 2½-meter operation

Tuning Up. To place the exciter in operation, it is best first to disconnect the lead between R_{11} and R_{12}, and ground the free end of R_{11}. This allows the oscillator to operate without the stabilizing action. A crystal near 7000 kc. should be plugged into the crystal socket, and a pair of headphones or an amplifier and speaker connected to the monitoring output connection. Speaking into the microphone and simultaneously tuning C_1 and C_2, an oscillator frequency will be found which allows the signal to be heard in the phones or speaker. The signal should be tuned for maximum volume by adjusting C_2, and then peaked further by adjusting the primary trimmer on the discriminator transformer, IFT.

Next, R_{11} should be disconnected from ground and reconnected to R_{12}. A receiver equipped with a b.f.o. should now be used to check the signal from the oscillator. The receiver should be tuned to a frequency near the sum of the crystal and discriminator transformer frequencies. If the crystal is near 7000 kc. and the transformer is tuned to around 450

kc., the receiver should be tuned near 7450 kc. Now, while slowly changing the oscillator frequency by tuning C_1, and following the signal with the receiver, a frequency should be found where the oscillator suddenly "pulls in" and changes frequency only slightly for quite a variation in C_1, indicating that this is the frequency at which the stabilizing circuit takes hold. If the opposite effect takes place, and the stabilizing circuit "throws" the oscillator from one frequency to another when C_1 is varied, it is an indication that the discriminator voltage is of the wrong polarity for stabilization. The remedy for this effect is simply to reverse the polarity of the discriminator voltage by moving the ground connection to R_{13} from R_{12}, and connecting R_{11} and C_{23} to R_{13}, or to reverse the connection from the secondary of IFT to the 6H6 plates. The latter method will usually be found to be the simplest, and these leads should be left long enough when the exciter is wired to allow the change to be made, if necessary.

A check for the operation of the stabilizing circuit may be made by bringing a hand near the 6F6 grid coil. When the crystal is removed from the socket, the frequency will vary greatly when this coil is approached with the hand. With the crystal in the socket, however, only a very slight frequency variation should take place when the hand is brought near the coil.

The oscillator output frequency will, as explained above, depend on the crystal and discriminator transformer frequencies. Since the transformer specified can be tuned over a range from 255 to 550 kc. with its own trimmers, the output frequency may be set anywhere from 7255 to 7550 kc. by simply readjusting the transformer, when a 7000-kc. crystal is used. Frequencies between 7316 and 7500 kc. are the only ones which should be used for the 10- and 5-meter f.m. bands, however. To use the exciter with a $2\frac{1}{2}$-meter transmitter, the output frequency should be between 7000 and 7250 kc. In this case, the crystal should have a frequency close to 7500 kc., allowing the discriminator to operate on the difference instead of the sum frequency. Changing from difference to sum frequency operation will require that the discriminator output voltage be reversed, as described above. If the change is to be

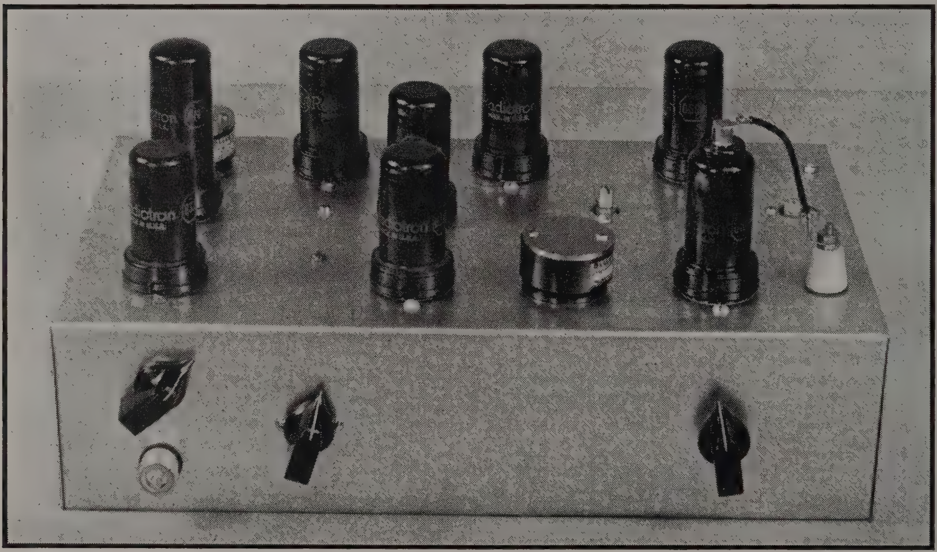


Figure 24.
PHASE-MODULATED EXCITER.

This chassis contains all the essentials for producing a crystal-controlled, phase-modulated f.m. signal, plus a circuit for increasing the deviation. The speech amplifier tubes are at the left edge of the chassis, with the low-frequency crystal behind the rear speech tube. Along the rear edge of the chassis are the crystal oscillator tube, the mixer, and a tripler stage. The converter tube and the high-frequency crystal are at the right front of the chassis, while the two balanced-modulator tubes are located near the center of the chassis. Controls on the front of the chassis are, left to right: gain control; low-frequency oscillator tuning; converter plate tuning.

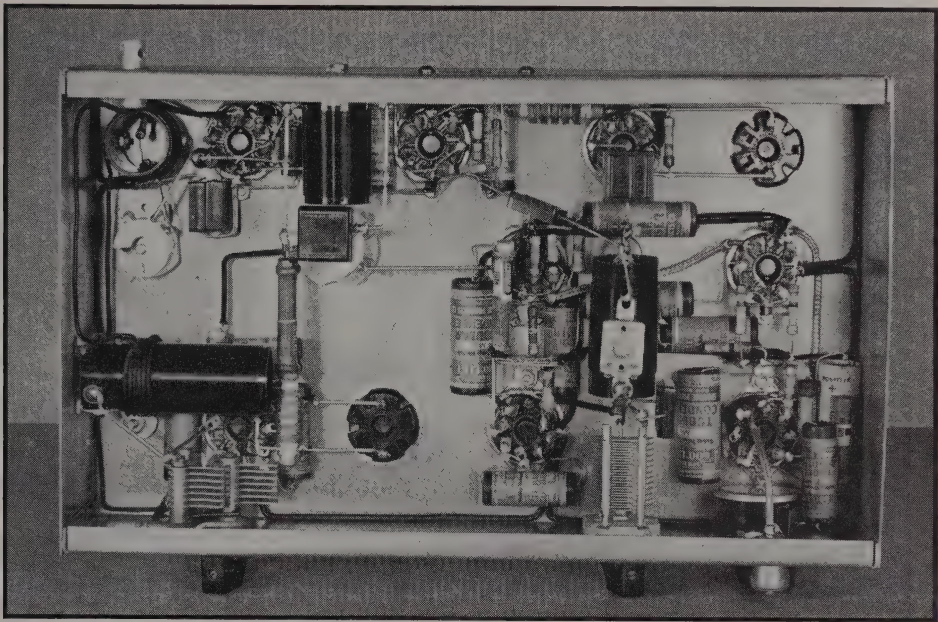


Figure 25.
P.M. EXCITER—BOTTOM VIEW.
Small receiving-type components are used throughout the exciter. The location of the various parts is discussed in the text. Note the auto-type link-output connector at the rear of the chassis. A 4-prong socket, also at the rear of the chassis, serves for power connections to the unit. Shielded wire is used for plate and grid leads in the audio section.

made often, a switch should be incorporated for that purpose.

The speech amplifier has sufficient gain to give satisfactory operation with any of the ordinary crystal or dynamic microphones. A deviation of 25 kc. on 10 meters may be easily obtained, with proportionally higher deviation on the higher frequencies. The gain control may be calibrated in terms of deviation by the method described in Chapter 9.

Phase-Modulated F.M. Exciter

By means of phase modulation it is possible to obtain a crystal-controlled f.m. signal. To obtain the phase-modulation, a constant frequency signal is amplitude modulated, and the a.m. sidebands are shifted 90 degrees in phase and recombined with the carrier. The basic theory of the phase-modulation scheme is discussed in detail in Chapter 9.

A phase-modulated f.m. exciter which gives output in or near the 160-meter amateur band suitable for frequency multiplication into the 10-, 5-, or 2½-meter f.m. bands is shown in Figures 24 and 25 and diagrammed in Figure 26, on the following page.

Circuit. The circuit consists basically of a 6J5 crystal oscillator on 1800 kc., a balanced modulator using a pair of 6SA7's, a 6SJ7 mixer on 1800 kc., another 6SJ7 as a tripler to 5400 kc., and a 6K8 crystal-controlled converter to bring the signal back to the 160-meter band. The speech amplifier uses a 6SJ7 followed by a 6N7 self-balancing phase inverter.

The circuit operates as follows: The output across the balanced 6J5 plate circuit is applied to the "oscillator" grids of the two 6SA7's. The latter tubes have their plates connected in parallel and operate into a common plate tank with the 6SJ7 mixer tube. Due to the fact that the plates of the 6SJ7's are in parallel, while their grids are in push-pull, there is no output from these tubes into the common plate circuit until they are unbalanced by application of audio voltage to their "signal" grids. When the audio modulation is applied, the 6SJ7's produce amplitude modulation sidebands, minus the carrier, in their plate circuit.

By means of the phase-shifting network, C_{10}-R_4, across L_1, the excitation to the 6SJ7 mixer stage is shifted 90 degrees with respect to the excitation on the 6SA7's. When the a.m. sidebands from the 6SA7's are combined

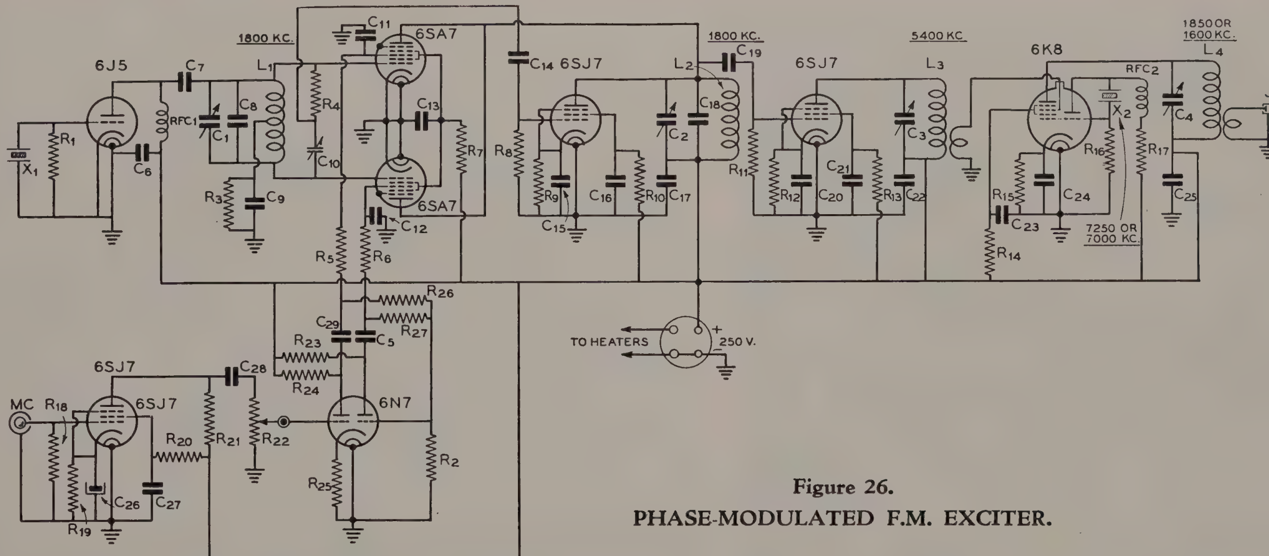


Figure 26.

PHASE-MODULATED F.M. EXCITER.

C_1, C_2—75-μμfd. midget variable
C_3—50-μμfd. midget variable
C_4—100-μμfd. midget variable
C_5—.05-μfd. 600-volt tubular
C_6, C_7—.01-μfd. 600-volt tubular
C_8—.0001-μfd. mica
C_9—.05-μfd. 600-volt tubular
C_{10}—3-30-μμfd. mica trimmer
C_{11}, C_{12}—.01-μfd. 600-volt tubular
C_{13}—0.1-μfd. 600-volt tubular
C_{14}—.003-μfd mica
C_{15}, C_{16}, C_{17}—.01-μfd. 600-volt tubular
C_{18}—.0001-μfd. mica
C_{19}—.0005-μfd. mica
C_{20}, C_{21}, C_{22}, C_{23}—.01-μfd. 600-volt tubular
C_{24}—.003-μfd. mica
C_{25}—.01-μfd. 600-volt tubular
C_{26}—10-μfd. 25-volt electrolytic
C_{27}—0.1-μfd. 600-volt tubular
C_{28}—.01-μfd. 600-volt tubular
C_{29}—.05-μfd. 600-volt tubular
R_1, R_2—100,000 ohms, ½ watt
R_3—10,000 ohms, ½ watt
R_4—5000 ohms, ½ watt
R_5, R_6—3000 ohms, ½ watt
R_7—10,000 ohms, 5 watts
R_8—100,000 ohms, ½ watt
R_9—300 ohms, ½ watt
R_{10}—50,000 ohms, ½ watt
R_{11}—100,000 ohms, ½ watt
R_{12}—300 ohms, ½ watt
R_{13}, R_{14} — 50,000 ohms, ½ watt
R_{15}—300 ohms, ½ watt
R_{16}, R_{17} — 50,000 ohms, ½ watt
R_{18} — 2 megohms, ½ watt
R_{19}—500 ohms, ½ watt
R_{20}—1 megohm, ½ watt
R_{21}—250,000 ohms, ½ watt
R_{22}—500,000-ohm potentiometer
R_{23}, R_{24}—100,000 ohms, ½ watt
R_{25}—2000 ohms, ½ watt
R_{26}, R_{27}—250,000 ohms, ½ watt
RFC_1, RFC_2—2½ mhy.
MC—Microphone connector
J—Auto-type connector
L_1, L_2—45 turns of no. 24 enam. close-wound on 1" dia. form. L_1 center-tapped
L_3—28 turns of no. 24 enam. close-wound on 1" dia. form. Secondary is 6 turns of hook up wire wound over "cold" end.
L_4—100 turns of no. 24 enam. close-wound on 1" dia. form. Link is 6 turns of hook up wire wound over "cold" end.
X_1—Crystal near 1800 kc.
X_2—Crystal near 7250 or 7000 kc., see text

in the plate of the mixer with the steady, phase-shifted carrier from the crystal stage, the result is a phase-modulated signal. To convert the phase modulation to frequency modulation, the phase modulation is made to decrease linearly with frequency by the R-C networks, R_5-C_{11} and R_6-C_{12}.

The section of the exciter following the 6SJ7 mixer stage is simply for the purpose of increasing the f.m. deviation to useable amounts. The first stage following the mixer is a 6SJ7 which triples the frequency to 5400 kc. Output from the tripler is applied to the 6K8 converter, which uses a Pierce crystal oscillator circuit in its oscillator section. When a 7250-kc. crystal is used in the 6K8, the difference frequency obtained in the converter plate circuit, L_4-C_4, is 7250 minus 5400, or 1850 kc. When the signal on 1850 kc. is multiplied 16 or 32 times in frequency by a series of doublers and/or quadruplers, the 10- and 5-meter f.m. bands are reached. It is also possible to use an additional doubler stage to reach the 2½-meter band, if desired. However, it is usually preferable to use a push-pull tripler from 7½ meters (38 Mc.) to excite a 2½-meter final amplifier, and for this reason provision is made in the exciter to reach 38 Mc. with the proper combination of frequency multipliers.

To give output from the exciter suitable for frequency multiplication to 38 Mc., a 7000-kc. crystal is used in the oscillator section of the 6K8. This gives a difference frequency of 1600 kc. in the converter plate circuit. A frequency multiplication of 24 times then gives a frequency of 38.4 Mc., which is suitable for tripling into the 2½-meter band. The frequency multiplication of 24 times may be obtained from one tripler and three doublers, or a tripler, a quadrupler—and a doubler. The required frequency multiplier stages may be arranged in any desired order, of course.

Construction. The exciter is constructed on a 7 x 12 x 3-inch chassis, which allows ample room for the components without crowding. The r.f. section of the exciter starts at the rear left corner of the chassis with the 1800-kc. crystal. The 6J5 is directly to the right of the crystal, followed to its right by the 6SJ7 mixer and the 6SJ7 tripler. The two 6SA7's are located side by side near the center of the chassis, and on a line passing between the oscillator and mixer tubes. At the right front corner of the chassis is the 6K8 converter, with its crystal to the left of the tube. The two speech amplifier tubes are located near the left edge of the chassis, progressing from front to rear.

Most of the under-chassis components are visible in Figure 25. In the photo, the speech amplifier stages are at the right, with the crystal socket at the right top. The r.f. section of the exciter runs along the top of the photo, ending up with the 6K8 stage at the lower left corner. The crystal stage plate coil is alongside the 6SA7's, which it feeds directly, at the right-center of the photo. The tank condenser across this coil, C_1, is mounted on the front drop of the chassis, but it could just as well be mounted vertically with the shaft extending through the chassis top, since the condenser setting need not be changed after the initial tuning. C_1 and C_2, the mixer and tripler tank condensers, are mounted with their shafts projecting through the top surface of the chassis, since they, too, are set once and then left alone. The converter plate-tank condenser, C_4, is located on the front drop of the chassis and fitted with a knob, since this circuit must be retuned if different crystals are used at X_2 when changing from 5- and 10-meter operation to 2½-meter operation.

Operation. To place the exciter into operation, it should be connected to a power supply delivering 250 to 300 volts at 80 to 100 ma., and 6.3 volts at 3 amperes. Crystals near the correct frequencies should be placed in the crystal sockets, and the two 6SA7's removed from their sockets. The r.f. section of the exciter may now be tuned up. The tuning procedure up through the plate of the tripler stage is no different from any ordinary multi-stage r.f. unit—simply tune each stage for maximum output as indicated by a flashlight lamp and loop or a neon bulb. If the coil data are followed accurately, L_3-C_3 cannot be tuned to any harmonic of the crystal other than the third, so it will not be necessary to check the frequency of this circuit with a wavemeter.

If the 6K8 crystal oscillator section is operating properly, it is only necessary to connect a flashlamp bulb across the link output terminals at J and tune C_4 for resonance at the difference frequency to complete the major part of the tuning-up operation. The oscillator performed satisfactorily with all of the crystals tried in the laboratory, so there should be no trouble from that source.

After the r.f. section has been tuned up, the 6SA7's may be replaced in their sockets, and C_1 reset slightly, if necessary, to restore the crystal to oscillation. Since the 6SA7's get their bias from the grid leak, R_3, they should not have plate voltage applied any longer than necessary when they are not being excited from the crystal stage. When the modulator tubes are in their sockets it will be necessary to retune C_2 slightly, decreasing its capacity to make up for the output capacity of the tubes which has been added across the circuit.

For a 90-degree phase shift in the excitation to the mixer stage, C_{10} should be set so that its reactance at the operating frequency (1800

kc.) is equal to the resistance of R_4 (5000 ohms). The correct capacity for C_{10} is thus 18-$\mu\mu$fd. The setting is not extremely critical, however, and it only is necessary to turn the adjusting screw on the condenser until the movable plate is about two-thirds of the way toward the maximum-capacity position.

Next, the microphone may be connected and the gain control turned up. Speaking into the microphone should now produce a signal which, when heard on an ordinary a.m. receiver with the b.f.o. operating, sounds like it is "splattering" terribly. The "splatter" should be much worse on the converter output frequency (1850 or 1600 kc.) than on the crystal frequency (1800 kc.), showing that the deviation is being increased by the frequency multiplier. If the receiver's b.f.o. is turned off and the receiver tuned slightly to one side of the signal, the modulation should come through satisfactorily. When the receiver is tuned exactly to resonance with the signal, it should be impossible to hear more than just a slight trace of modulation.

Deviation. The maximum amount of f.m. deviation produced by the exciter will depend upon the transmitter output frequency and the setting of the gain control. For speech work, the deviation at the plate of the 6SJ7 mixer stage may be as high as 200 cycles, without appreciable distortion. This deviation is increased three times in the exciter itself, so that the exciter output deviation can be 600 cycles without difficulty. Frequency multiplication of 16 times from the exciter to the 10-meter band gives a deviation capability of almost 10 kc. On the 5-meter band the deviation will be twice this amount, or 20 kc. To reach the 2½-meter band from the 1600-kc. exciter output frequency requires a frequency multiplication of 72 times, and this gives a maximum deviation, under the above conditions, of approximately 43 kc. The total "swing" will be twice the above figures, or 20 kc. on 10 meters, 40 kc. on 5 meters, and 86 kc. on 2½ meters.

More or less deviation than in the examples just illustrated may be obtained simply by increasing or decreasing the gain in the speech amplifier. Higher gain will increase the deviation with an accompanying increase in distortion at the lower audio frequencies, and less gain will reduce the deviation with a decrease in distortion at the lower audio frequencies. The "carrier null" method described in Chapter 9 may be used to measure the actual amount of deviation for any given setting of the gain control.

Crystal Frequencies. The crystal frequencies given in the preceding description need not be adhered to exactly, of course. The only requirement is that the output frequency of the exciter must fall between 1829 and 1875 kc. for operation in the 5- and 10-meter f.m. bands, and between 1556 and 1611 kc. for multiplication via a tripler into the 2½-meter band. Crystals on or near the frequencies shown in the diagram are in amateur bands, however, and are therefore less expensive than those ground to special frequencies.

50-Watt F.M. Transmitter

The transmitter illustrated in Figures 27 and 28 and diagrammed in Figure 29 has an output of 50 watts, frequency modulated, at 112 Mc. Except for the 500-volt, 250-ma. power supply, the transmitter is all located on the single 10 x 23 x 3-inch chassis.

The Exciter Stages. The exciter section of the transmitter, which includes those stages preceding the output stage, employs standard receiving tubes throughout, and is distinguished by the lack of unusual or "trick" circuit arrangements. The frequency modulated oscillator is a 6F6 operating at low plate and screen voltages to assure minimum frequency drift. The grid circuit of this stage is tuned to 9.5 Mc., and the plate circuit to the second harmonic, or 19 Mc.

Excitation from the oscillator stage is carried through a small coupling condenser to the grid of the following doubler, which is a 6V6-GT. To reduce the number of tuned circuits in the transmitter, the plate circuit of the doubler stage is tuned only by being closely inductively coupled to the grid circuit of the following stage. The close coupling required for proper operation of this type of circuit is achieved by locating the 6V6-GT plate coil, L_3, inside the following grid coil, L_4. There are no rigorous requirements in the choice of the tube used in the doubler stage, and a 6V6 or 6L6 may be substituted for the 6V6-GT, if desired, without making any circuit changes. If the plate voltage to the doubler stage is lowered somewhat, a 6F6 may be used. The doubler output frequency is 38 Mc.

Following the doubler stage is a tripler to 114 Mc. employing push-pull 7A4's. Circuit balance in this stage is provided by grounding the rotor of the split stator grid condenser. It is not necessary to ground the rotor of the plate tank condenser in this stage. No harm will be done, however, if the type of plate condenser used makes grounding the rotor more convenient than insulating it.

Tests with an experimental version of this transmitter showed the necessity of placing the tuned input circuit of the tripler directly in the grid circuit, rather than the more conventional method of capacity coupling from a balanced, tuned plate circuit in the preceding stage. With the latter type of circuit, the

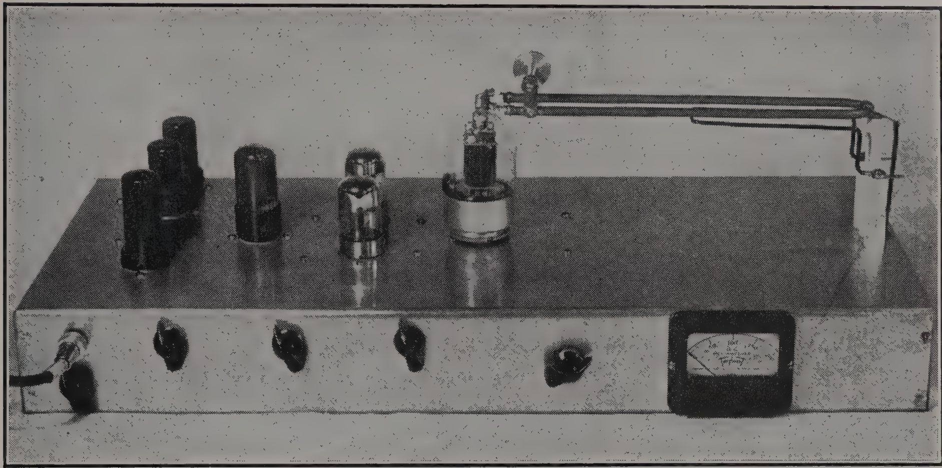


Figure 27.
TOP VIEW OF THE 50-WATT F.M. TRANSMITTER.
The three tubes in the row on the left edge of the chassis are, front to back: the 6F6 oscillator, 6SJ7 reactance tube, and 6SC7 speech amplifier. Then comes the 6V6-GT doubler to 38 Mc. and the push-pull 7A4 tripler to 112 Mc., followed by the 50-watt output tube and its tank circuit. Note the manner in which the plate circuit tubing condenser has been soldered to the plate rods.

tripler stage is prone to oscillate, while with the circuit shown in the diagram there is no tendency toward oscillation or instability.

By itself, the exciter section of the transmitter forms a complete, inexpensive, low power 112-Mc. transmitter with an output of 5 to 7 watts, and if the constructor is interested in a transmitter in this power class he could well choose the exciter of the 50-watt transmitter.

Output Stage. The 50-watt output stage utilizes an 815 tube, which has two beam tetrodes, each of somewhat lower power rating than an 807, in a single envelope. Excitation to the 815 is obtained by a 2-turn coupling coil pushed between the center turns of L_5. The excitation is adjusted by pushing the coupling coil in and out of L_5—too much coupling will overload the tripler and reduce the excitation and output in the final amplifier, while too little coupling will reduce the excitation and output. It is a simple matter to adjust the excitation properly by observing the output from the transmitter.

The final plate tank circuit consists of a U-shaped piece of $\frac{1}{4}$-inch copper tubing measuring $9\frac{1}{2}$ inches on each leg, with the two legs separated 1 inch. Tuning of the linear tank circuit is accomplished by varying the spacing between the plates of a small condenser at the plate ends of the tank circuit. The condenser plates and their supporting strips were taken from a small neutralizing condenser originally intended for neutralizing a 6L6. The plates and the supporting metal were removed from the insulator assembly which originally served as a mounting for the condenser, and the metal strips were soldered to the tank circuit with the aid of a small alcohol torch.

The antenna coupling "hairpin" is made up of a length of no. 10 enamelled wire supported by two stand-off insulators which also serve as terminals for connecting the antenna feeders. The antenna coupling is varied by bending the coupling hairpin toward or away from the plate tank.

Modulator and Speech Amplifier. The frequency modulator uses a conventional reactance tube circuit. The theory of operation of this type of circuit is described in Chapter 9. Partially fixed bias on the reactance tube is provided by resistor R_4, which bleeds a constant amount of current through the reactance tube bias resistor, R_3. Varying amounts of positive and negative d.c. voltage may be applied across the grid resistor, R_1, to determine whether the frequency varies linearly each side of the "carrier" frequency when the control voltage is varied. Non-linearity may be corrected by changing the value of R_4. In the transmitter shown, the resistor value specified in the diagram caption gave a linear voltage-frequency characteristic. For a 50-kc. swing under modulation at 112 Mc., the modulator should be linear over a range of slightly more than 4 kc. at the oscillator frequency.

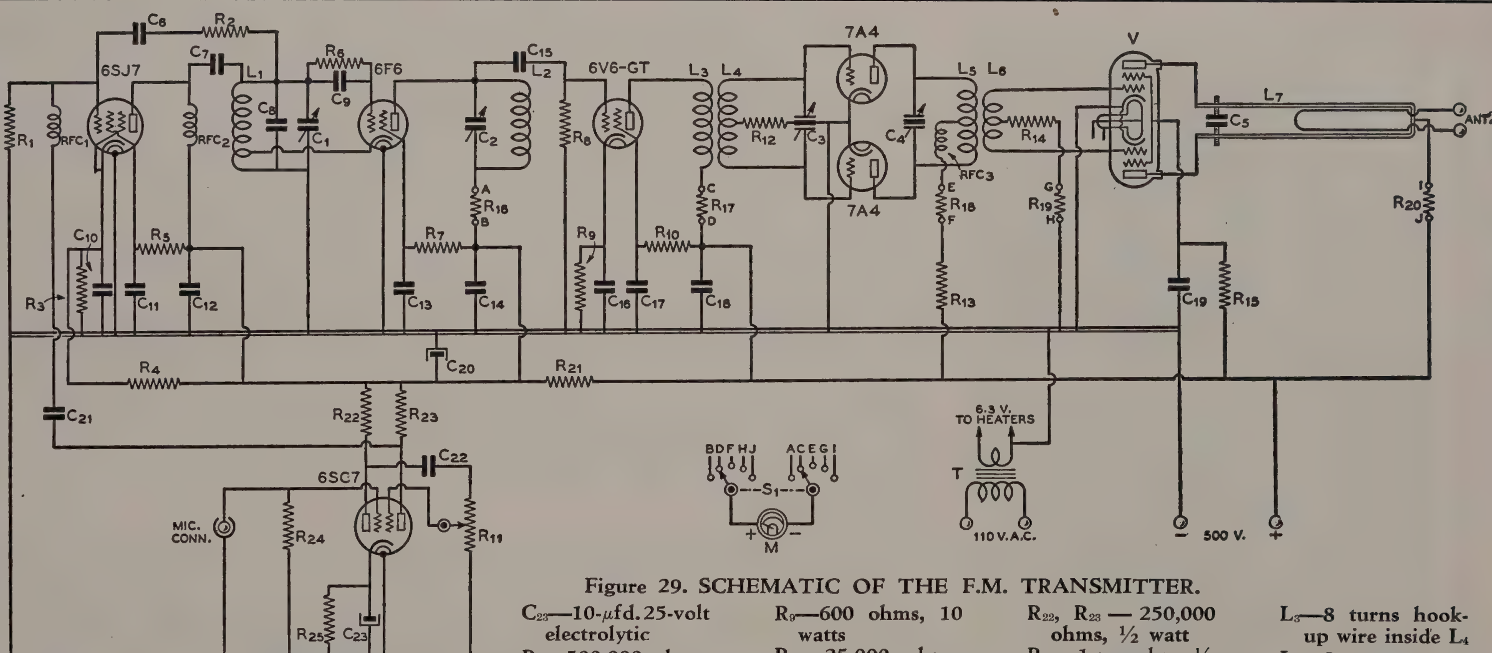


Figure 29. SCHEMATIC OF THE F.M. TRANSMITTER.

C_1, C_2—50-$\mu\mu$fd. midget variable
C_3—50-$\mu\mu$fd. per section
C_4—25-$\mu\mu$fd. per section
C_5—Midget circular plate neut. condenser mounted on plate rods
C_6, C_7—.003-μfd. midget mica
C_8—.0001-μfd. zero coeff.
C_9—.0001-μfd. midget mica
C_{10}, C_{11}, C_{12}—.003-μfd. mica
C_{13}, C_{14}—.003-μfd. mica
C_{15}—.00005-μfd. mica
C_{16}, C_{17}, C_{18}—.003-μfd. mica
C_{19}—.0003-μfd. silver mica
C_{20}—8-μfd. 450-volt electrolytic
C_{21}—.01-μfd. 400-volt tubular
C_{22}—.005-μfd. mica
C_{23}—10-μfd. 25-volt electrolytic
R_1—500,000 ohms. ½ watt
R_2—50,000 ohms. ½ watt
R_3—1000 ohms, ½ watt
R_4—60,000 ohms, 1 watt
R_5—100,000 ohms, ½ watt
R_6—50,000 ohms, ½ watt
R_7—25,000 ohms, 1 watt
R_8—100,000 ohms, 1 watt
R_9—600 ohms, 10 watts
R_{10}—25,000 ohms, 1½ watts
R_{11}—500,000 - ohm potentiometer
R_{12}—10,000 ohms, 1½ watts
R_{13}—5000 ohms, 10 watts
R_{14}—10,000 ohms, 1½ watts
R_{15}—25,000 ohms, 1½ watts
R_{16}, R_{17}, R_{18}, R_{19}, R_{20}—100 ohms, 1 watt
R_{21}—5000 ohms, 10 watts
R_{22}, R_{23} — 250,000 ohms, ½ watt
R_{24}—1 megohm, ½ watt
R_{25}—1000 ohms, ½ watt
RFC_1, RFC_2 — 2.5 mh. 125 ma.
RFC_3—U.h.f choke
T—6.3-v. 10-a.
M—0-200 d.c. milliammeter
L_1 — 12 t. no. 18 d.c.c. 1" dia. 1" long
L_2 — 13 t. no. 14 enam., ⅝" dia. 1" long
L_3—8 turns hookup wire inside L_4
L_4—8 turns no. 14 enam. ⅞" dia. 1¼" long
L_5 — 4 t. no. 14 enam. ⅝" dia. ¾" long
L_6—2 turns hookup wire wound with L_5
L_7—20" ¼" copper tubing bent into a hairpin with 1" center-to-center spacing
V—815 dual beam tetrode

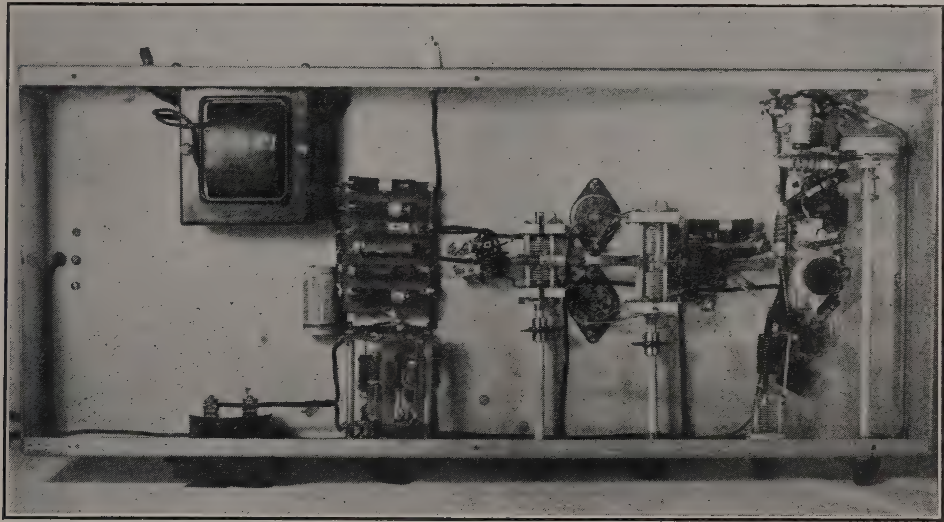


Figure 28.
UNDER-CHASSIS VIEW OF THE 50-WATT F.M. TRANSMITTER.
The filament transformer for the entire rig is shown in the left rear. The meter switch is mounted, with the resistors between its sections, directly in front of the resistor tie plate for the rig.

A single 6SC7 dual triode is used as a 2-stage speech amplifier. This tube provides considerably more voltage gain than is necessary to give a 50-kc. swing, when a crystal microphone is used. The 6SC7 may be replaced by a low gain triode (6C5, 6J5, etc.) if a low output single-button microphone or a double-button microphone is used. High output single-button microphones (telephone type) may be coupled directly into the reactance tube control grid by a microphone transformer, with the gain control, R_{11}, replacing the fixed grid resistor R_1.

Construction. As the photographs show, all of the wiring except the 815 plate circuit is below the chassis. The oscillator, modulator, and speech amplifier circuit occupy the space toward the left edge of the chassis. The stages following the oscillator are placed along the center line of the chassis, with each circuit placed as close to the preceding one as possible, since short leads are of prime importance in the high- and ultra-high-frequency stages. In each of the stages following the oscillator, all ground returns are brought, through separate leads, to a single point on the tube socket.

Operation. To place the transmitter into operation, 250 to 300 volts should be applied to the "+500" terminal and the oscillator first tuned to 9500 kc., as indicated by a conventional receiver. After the oscillator grid circuit has been set to the correct frequency, the oscillator plate circuit and following stages should each be tuned to resonance as indicated by minimum plate current. It will be found that tuning the oscillator plate circuit to the second harmonic of the grid circuit frequency will change the oscillator frequency slightly, and it may be necessary to retune the grid circuit after the plate circuit has been resonated. After the complete transmitter has been tuned up, the antenna may be connected and the full 500 volts applied.

Typical current readings, at resonance, are as follows: Oscillator plate—15 ma.; doubler plate—35 ma.; tripler plate—40 ma.; final amplifier grid—3 ma.; final amplifier plate—150 ma.

Although it is not to be recommended except for extremely short periods of time, a check on the operation of the output stage may be made by removing the loading and observing the minimum plate current. If the stage is operating correctly, the plate current will be approximately 30 ma. at resonance without load.

MICROWAVE TRANSMITTERS

Microwaves are generally considered as being those whose wavelength is less than 1 meter (frequencies greater than 300 Mc.). Microwaves are generated by means of *mag-*

netrons, electron-orbit oscillators, and *regenerative oscillators.* Microwaves are used by broadcast stations for remote pickup, by amateurs and experimenters, and for occasional telegraph and telephone communication, such as the British channel-spanning system. The technical problems encountered in this field are numerous, yet new tubes designed for microwaves have simplified many of these problems, and have been instrumental in increasing the usefulness of the band.

The Magnetron Oscillator. The magnetron is a specially designed tube for very-short-wave operation. It consists of a filament or cathode between a split plate, as shown in Figure 30.

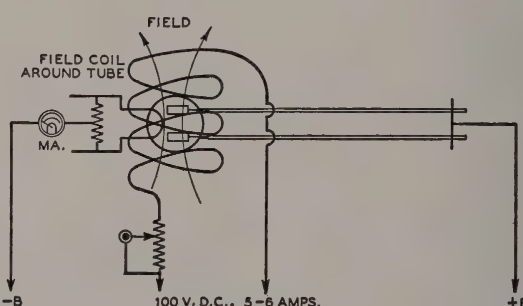
A magnetic field is produced at the filament by means of a large external field coil which is energized by several hundred watts of d.c. power. Ultra-high-frequency oscillations are produced in the split plate circuit when this magnetic field is in the correct direction and of the proper intensity. A parallel-wire tuned circuit should be used for wavelengths below 1 meter. The frequency stability is not very good, and it is difficult to obtain satisfactory voice modulation from magnetron oscillators.

Electron-Orbit Oscillator. The range of oscillation in ordinary circuits is limited by the time required for electrons to travel from cathode to anode. This transit time is negligible at low frequencies, but becomes an important factor below 5 meters. With ordinary tubes, oscillation cannot be secured below 1 meter, but by means of *electron-orbit oscillators,* in which the grid is made positive and the plate is kept at zero or slightly negative potential, oscillation can be obtained on wavelengths very much below 1 meter.

Parallel-wire tuning circuits can be connected to these tube oscillators in order to increase the power output and efficiency. The tubes most suitable for this type of operation have cylindrical plates and grids, and their output is limited by the amount of power which can be dissipated by the grids. For transmitting, tubes such as the 35T, HK-54, 852, etc., can be used in the circuit shown in Figure 31, which is a modification of the circuit of Figure 32. More output is obtained by using a tuned-cathode circuit instead of tuned-grid circuit. Modulation can be applied to either the plate or grid. The frequency stability is very poor.

Regenerative Oscillators. The introduction of RCA "acorn" tubes made low power ½-meter regenerative oscillators practical. These tubes are more efficient than ordinary types for ultra-high-frequency work, and are available in several types in both 6.3 v. and

Figure 30.
SPLIT-ANODE MAGNETRON MICRO-WAVE OSCILLATOR.
Special magnetron tubes delivering several watts output at extremely high frequencies are available for certain experimental purposes. Their main disadvantage for amateur work is that they are rather difficult to obtain. Also, a source of d.c. of large magnitude is required for the field electromagnet.



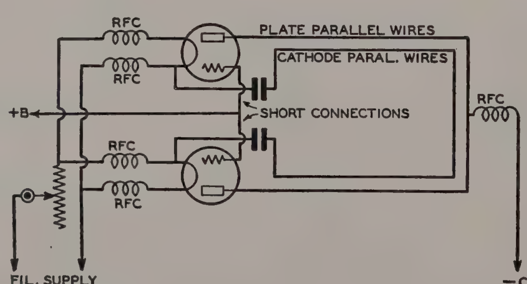


Figure 31.
KOZANOWSKI OSCILLATOR.
This type of u.h.f. oscillator requires the use of tubes having cylindrical elements, such as the HK-24 and 54, 35TG and 75T, 1628 and 852, and HY-75. Certain receiving tubes such as the 7A4, 955, etc. may also be used for lower power outputs. The grid dissipation of the tubes is the most important limiting factor on their power output.

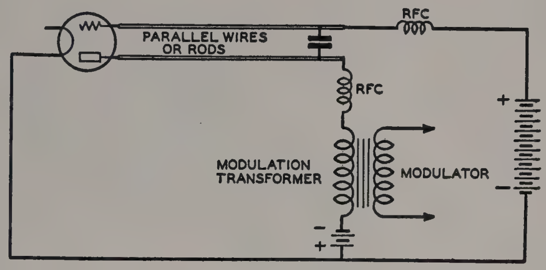


Figure 32.
BARKHAUSEN-KURTZ OR GILL-MORRELL OSCILLATOR.
As with all oscillators of the electron-orbit type, the grid dissipation will be very high and the oscillator tube should have cylindrical elements. This circuit may also be used as a detector for micro-wave signals, with the transformer feeding into an audio amplifier instead of being fed audio energy for modulation.

1.4 v. series. They are satisfactory for low-power transmitters and superregenerative receivers. The regenerative circuits are quite similar to those for longer wavelengths, except for the physical size of condensers and coils. The tube element spacing in these acorn tubes is made so small that electron transit time becomes a negligible factor for wavelengths above 0.6 meter.

Acorn tubes are also made in r.f. pentode amplifier types, both sharp cutoff and remote cutoff. However, these require concentric tank circuits below 2½ meters, because at such high frequencies it is impossible, due to high losses, to obtain appreciable gain (high Q) with conventional tanks.

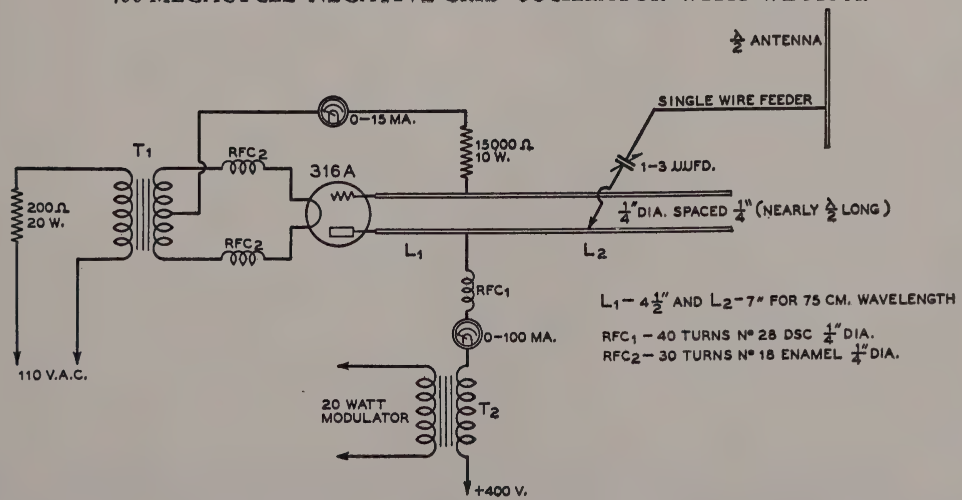
For higher power oscillators, special transmitting tubes designed for microwave work are offered by several manufacturers, notably Western Electric, Hytron, RCA, and Eimac. The HK-24 also makes an excellent microwave tube when two are used in push-pull.

For maximum output at 2½ meters and shorter wavelengths, filament chokes are sometimes required. One way to avoid the necessity for filament chokes and at the same time increase the efficiency is to substitute a tuned filament circuit for the usual tuned grid circuit, by-passing the grids to ground.

Microwave regenerative oscillators are most efficient when linear tank circuits are used in place of coils, and when two tubes are used in push-pull. Maximum output and efficiency cannot be obtained with single-ended circuits.

¾-Meter Parallel Rod WE-316-A Transmitter. A large variety of circuits could be suggested for microwave operation, but the most simple of these is the one shown in Figures 33 and 34. It consists of two parallel half-wave rods, spaced about ¼ inch apart,

Figure 33.
400-MEGACYCLE NEGATIVE-GRID OSCILLATOR WITH WE-316-A.



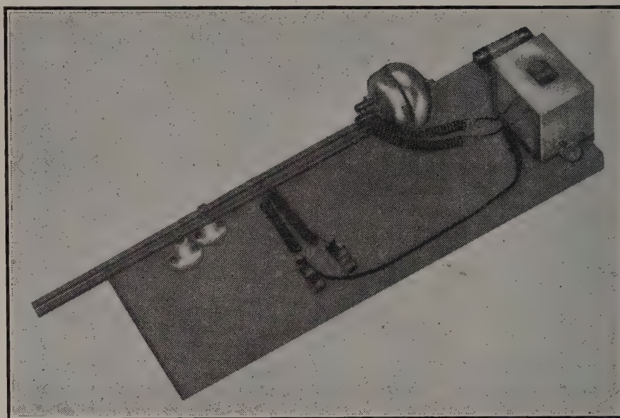


Figure 34.
WE-316-A 400-MEGACYCLE OSCILLATOR.

to provide a $\frac{3}{4}$-meter tuned circuit of fairly high Q. The grid and plate of the tube are connected to the copper rods; this capacity causes the physical length to be less than a half wavelength. As can be seen from the photograph, the plate r.f. choke and the grid leak do not connect to the center of the rods, but rather across the voltage node. The distance between this point and the free ends of the rods is a quarter wavelength.

Filament r.f. chokes, or tuned filament leads, are desirable for operation below 1 meter because the filament is not strictly at a point of ground potential in the oscillating

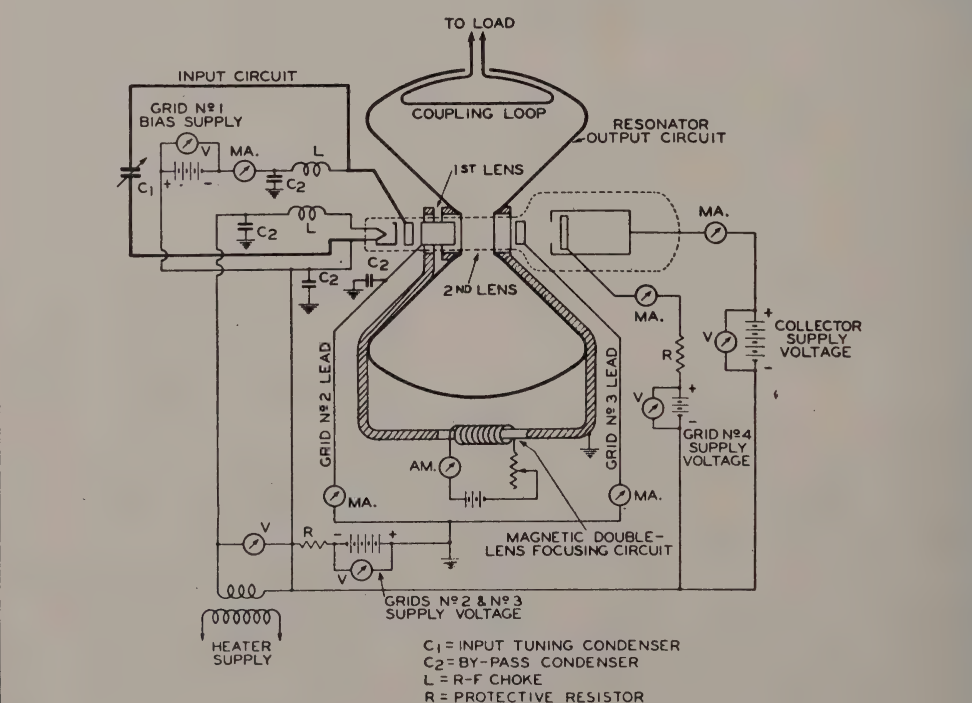


Figure 35.
OPERATING CIRCUIT FOR THE 825 AS A 500-MEGACYCLE AMPLIFIER.

circuit. These filament chokes consist of 30 turns of no. 16 enamelled wire, wound on a $\frac{1}{4}$-inch rod, then removed from the rod and air-supported, as the picture shows. The length of these chokes is approximately 3 inches. A 200-ohm resistor is placed in series with the 110-volt a.c. line to the filament transformer in order to reduce the transformer secondary voltage from $2\frac{1}{2}$ to 2 volts, because the filament of the tube operates on 2 volts at 3.65 amperes. This particular oscillator gave outputs in excess of 5 watts on $\frac{3}{4}$ meter, even when no filament r.f. chokes were used.

Operation of the Oscillator. This oscillator, when loaded by an antenna, draws from 70 to 80 milliamperes at 400 volts plate supply. The oscillator should be tested at reduced plate voltage, preferably by means of a 1000- to 2000-ohm resistor in series with the positive B lead, until oscillation has been checked. A flashlight globe and loop of wire can be coupled to the parallel rods at a point near the voltage node, in order to indicate oscillation. A thermo-galvanometer coupled to a loop of wire makes a more sensitive indicator, but the high cost of this meter prohibits its use in most cases.

A 15-inch antenna rod or wire can be fed by a 1- or 2-wire feeder of the nonresonant type. A single-wire feeder can be capacitively coupled to the plate rod, either side of the voltage node, through a small blocking condenser. If a 2-wire feeder is employed, a small coupling loop, placed parallel to the oscillator rods with the closed end of the loop near the voltage node of the oscillator, will provide a satisfactory means of coupling to the antenna. U.h.f. antennas are described in Chapter 21.

Microwave Amplifiers

It is extremely difficult to get into operation any type of amplifier circuit of the conventional type at a frequency greater than about 250 Mc. The main reasons for this difficulty have been the extremely high amounts of loading of the interelectrode capacitances, the high inductances of leads to elements, and the practical impossibility of obtaining a satisfactory neutralizing arrangement. Quite recently, however, RCA announced an entirely new type of amplifier arrangement which had been under development for quite a period of time. In the new circuit arrangement, the output tuned circuit is inductively coupled to a stream of high-velocity electrons within a vacuum tube especially designed for the purpose. Since there is no coupling other than the purely inductive coupling between the electron stream and the output tank circuit, all the difficulties mentioned above have been averted.

A new vacuum tube especially designed for this service, the 825, has been placed upon the market. It is called an *inductive output amplifier* and is suitable for operation on frequencies above 300 Mc. Figure 35 shows a diagram of an inductive amplifier stage which is capable of delivering an output of 35 watts at a frequency of 500 megacycles.

CHAPTER TWENTY

Antenna Theory and Operation

RADIO waves consist of condensations and rarefactions of energy traveling through space with the speed of light, (186,000 miles or 300,000,000 meters per second). These waves have an electrostatic and an electromagnetic component. The electrostatic component may be considered as corresponding to the voltage of the wave and the electromagnetic component to the wave current. Radio waves not only travel with the speed of light but can be refracted and reflected much the same as light waves.

Radio Waves and Their Propagation

Polarization. Like light waves, radio waves can have a definite polarization. In fact, while light waves ordinarily have to be reflected or passed through a polarizing medium before they have a definite polarization, a radio wave leaving a simple radiator will have a definite polarization, the polarization being indicated by the orientation of the electrostatic component of the wave. This, in turn, is determined by the orientation of the radiator itself, as the electromagnetic component is always at right angles to a linear radiator, and the electrostatic component is always in the same plane as the radiator. Thus we see that an antenna that is vertical with respect to the earth will transmit a vertically polarized wave, as the electrostatic lines of force will be vertical. Likewise, a simple horizontal antenna will radiate horizontally polarized waves.

Because the orientation of a simple linear radiator is the same as the polarization of the waves emitted by it, the radiator itself is referred to as being either vertically or horizontally polarized. Thus, we say that a horizontal antenna is horizontally polarized.

The Ionosphere. A simple transmitting antenna or radiating system sends out radio waves in nearly all directions, though the strength of the waves may be greater in certain directions, and at certain angles above the earth. High frequency energy radiated along the surface of the earth is rapidly attenuated, and is of little use for consistent communication over distances exceeding 50 or 75 miles. That part of the radiated energy which is sent up at an angle above the horizon is partly returned to earth by the bending effect produced by the varying density of the ionized particles in the various layers of the *ionosphere*.

The ionosphere consists of layers of ionized particles of gas located above the stratosphere, and extending up to possibly 750 miles above the earth. Thus we see that high-frequency radio waves may travel over short distances in a direct line from the transmitter to the receiver, or they can be radiated upward into the ionosphere to be bent downward in an indirect ray, returning to earth at considerable distance from the transmitter. The wave reaching a receiver via the ionosphere route is termed a *sky wave.* The wave reaching a receiver by traveling in a direct line from the transmitting antenna to the receiving antenna is commonly called a *ground wave,* or *surface wave.*

The amount of bending which the sky wave undergoes depends upon the frequency of the wave, and the amount of *ionization* in the ionosphere, which is in turn dependent upon radiation from the sun. The sun increases the density of the ionosphere layers, and lowers their effective height. For this reason, radio waves act very differently at different times of day, and at different times of the year.

The higher the frequency of a radio wave, the farther it penetrates the ionosphere, and the less it tends to be bent back toward the earth. The lower the frequency, the more easily the waves are bent, and the less they penetrate the ionosphere. 160-meter and 80-meter signals will usually be bent back to earth even when sent almost straight up, and may be considered as being *reflected* rather than *refracted.* As the frequency is raised beyond about 5,000 kc. (dependent upon the critical frequency of the ionosphere at the moment), it is found that waves transmitted at angles higher than a certain critical

angle *never return to earth.* Thus, on the higher frequencies, it is usually desirable to confine radiation to low angles, since the high angle waves simply penetrate the ionosphere and keep right on going, never returning to earth.

Signals above about 45,000 kc. are bent so slightly that they seldom return to earth, regardless of the vertical angle of radiation, although, under exceptional circumstances, radio waves of 75,000 kc. have been known to return to earth for very short periods of time. Thus, sky wave propagation does not permit *consistent* communication at frequencies of 45,000 kc. In fact, the results on frequencies above 22,000 kc. are not considered consistent enough for commercial use.

Skip Distance. The ground wave of a 14,000-kc. transmitter can seldom be heard over 100 miles away. Also, the first bending of the sky wave rarely brings it back down to earth within 300 miles from the 14,000-kc. transmitting antenna at night. Thus, there is an area, including all distances between 100 and 300 miles from the transmitter, in which the signals are not ordinarily heard. The closest distance at which sky waves return to earth is called the *skip distance.* In the skip zone, no reception is possible, but moving closer to or farther away from the transmitter allows the signals to be heard.

Fading. The lower the angle of radiation of the wave, with respect to the horizon, the farther away will the wave return to earth, and the greater the skip distance. The wave can be reflected back up into the ionosphere by the earth, and then be reflected back down again, causing a second skip distance area. The drawing of Figure 1 shows the multiple reflections possible. When the receiver receives signals which have traveled over more than one path between transmitter and receiver, the signal impulses will not all arrive at the same instant, as they do not all travel the same distance. When two or more signals arrive in the same phase at the receiving antenna, the resulting signal in the receiver will be quite loud. On the other hand, if the signals arrive 180° out of phase, so they tend to neutralize each other, the received signal will drop,—perhaps to zero, if perfect neutralization occurs. This explains why high-frequency signals fade in and out.

Fading can be greatly reduced on the high frequencies by using a transmitting antenna with sharp vertical directivity, thus cutting down the number of multiple paths of signal arrival. A receiving antenna with similar characteristics (sharp vertical directivity) will further reduce fading. It is desirable, when

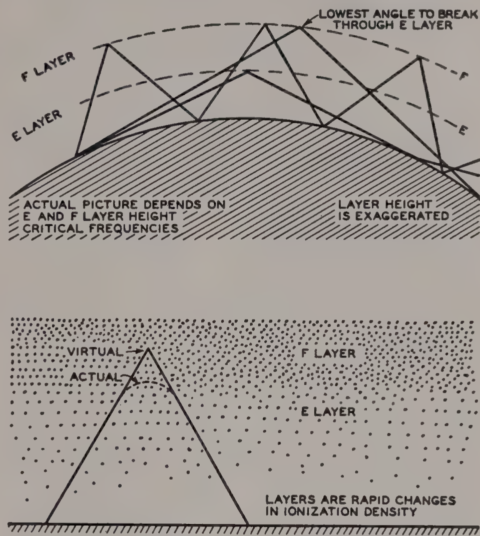


Figure 1.
Illustrating how the ionized atmosphere or ionosphere layer can bend radio waves back to earth, and some of the many possible paths of a high-frequency sky wave signal.

using antennas with sharp vertical directivity, to use the lowest vertical angle consistent with good signal strength for the frequency used. This cuts down the number of hops the signal has to make to reach the receiver, and consequently reduces the chance for arrival via different paths.

Selective Fading. Selective fading affects all modulated signals. A modulated signal is not a single frequency signal, but consists of a narrow band of waves perhaps 15 kc. wide. It will be seen that the whole modulated signal band may not be neutralized at any instant, but only part of it. Likewise, most of the carrier may be suppressed, or one sideband may be attenuated more than the other. This causes a peculiar and changing form of audio distortion at the receiver, which is known as *selective fading.*

Angle of Radiation. For a certain frequency, ionosphere height, and transmitting distance there is an optimum angle with the horizon at which the radio wave should be propagated. For extremely long distance communication, the angle of radiation should be low (5 to 15 degrees above the horizon), regardless of the frequency used, so that the wave may arrive in the fewest possible jumps. For comparatively short distance communication (between 100 and 400 miles), the optimum angle of radiation will be considerably higher, but because very high fre-

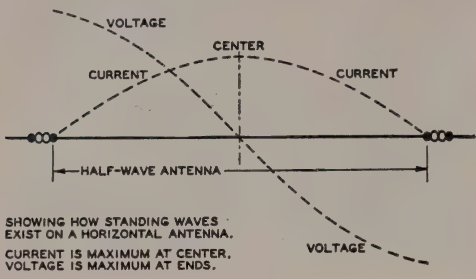


Figure 2.

quency waves are not readily bent, and penetrate the ionosphere when striking it at too steep an angle, we see that the shorter wavelengths are not satisfactory for short distance communication. Thus, we have the skip distance, or zone of silence, previously referred to. Different types of antennas have different major angles of radiation with respect to the earth and the antenna, as will be shown later.

Antenna Radiation

Alternating current passing through a conductor creates an alternating electromagnetic field around that conductor. Energy is alternately stored in the field, and then returned to the conductor. As the frequency is raised, more and more of the energy does not return to the conductor, but instead is radiated off into space in the form of electromagnetic waves, called radio waves. Radiation from a wire, or wires, is materially increased whenever there is a sudden *change* in the *electrical constants* of the line. These sudden changes produce reflection, which places *standing waves* on the line.

When a wire in space is fed radio frequency energy having a wavelength of approximately 2.08 times the length of the wire in meters, it *resonates* as a *dipole* or half-wave antenna at that wavelength or frequency. The greatest possible change in the electrical constants of a line is that which occurs at the open end of a wire. Therefore, a dipole has a great mismatch at each end, producing a high degree of reflection. We say that the dipole is terminated in an infinite impedance (open circuit). An incident radio frequency wave traveling to one end of the dipole is reflected right back towards the center of the dipole after reaching the end, as there is no place else for it to go.

A returning wave which has been reflected meets the next incident wave, and the voltage and current at any point along the antenna are the algebraic sum of the two waves. At the ends of the dipole, the voltages add up, while the currents of the two waves cancel, thus producing *high voltage* and *low current* at the *ends* of the dipole or half-wave section of wire. In the same manner, it is found that the currents add up while the voltages cancel at the center of the dipole. Thus, at the *center* there is *high current* but *low voltage.*

Inspection of Figure 2 will show that the current in a dipole decreases sinusoidally towards either end, while the voltage similarly increases. The voltages at the two ends of the antenna are 180° out of phase, which means that the polarities are opposite, one being plus while the other is minus at any instant. A curve representing either the voltage or current on a dipole represents a *standing wave* on the wire. If the voltage or current measured the same all along the wire, it would indicate the absence of standing waves. The latter condition can exist only when energy is absorbed from one end of a wire or line exactly at the same rate it is supplied to the other end. The latter condition is covered thoroughly later in the chapter, under the heading of *Untuned Transmission Lines.* Many transmission lines do not have uniform voltage and current along their length, and thus have standing waves the same as a dipole or antenna radiator.

A point of maximum current on a radiator or tuned resonant transmission line ordinarily corresponds to a point of minimum voltage. A *loop* means a point of *maximum* current or voltage, while a *node* refers to a point of *zero* or *minimum* current or voltage. Thus, we see that a voltage loop corresponds to a current node, and vice versa. In a wire or line containing reactance, this is not strictly true, but both antennas and tuned transmission lines ordinarily are operated at resonance, and the reactance, therefore, is negligible.

A 2-wire resonant line does not radiate appreciably in spite of its high reflection and consequent standing waves, because the radiation from the 2 adjacent wires is of opposite polarity or phase, and equal in amplitude, thus cancelling out if the spacing is but a very small fraction of a wavelength. In other words, the radiation from one wire is neutralized by the other wire, and vice versa.

Frequency and Antenna Length. All antennas commonly used by amateurs, excepting the terminated rhombic, are based on the fundamental Hertz type, which is a wire in space a half wavelength long electrically. A linear, resonant dipole, which is a half wavelength long *electrically,* is actually slightly less than a half wave long *physically,* due to the capacity to ground, "end effects", and the fact that the velocity of a high-frequency radio

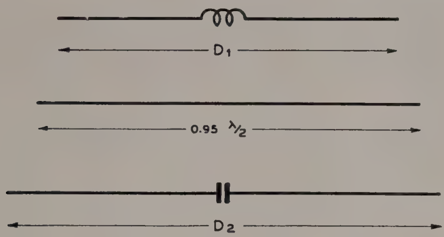


Figure 3.
THREE ANTENNAS ALL EQUAL ELECTRICALLY TO ONE HALF WAVELENGTH.
The top antenna is inductively lengthened. The bottom one is capacitively shortened. A coil will have the most lengthening effect and a condenser the most shortening effect when located at a current loop.

wave traveling along the conductor is not quite as high as it is in free space.

If the cross section of the conductor is kept small compared to a half-wavelength, these effects are relatively constant, so that an electrical half wave is a fixed percentage shorter than a physical half-wavelength. This percentage is approximately 5 per cent. Therefore, most linear half-wave antennas are close to 95 per cent of a half wave long physically. Thus, a half-wave antenna resonant at exactly 80 meters would be one-half of 0.95 times 80 meters in length. Another way of saying the same thing is that a wire resonates at a wavelength of about 2.1 times its length in meters. If the diameter of the conductor begins to be an appreciable fraction of a wavelength, as when copper tubing is used as an u.h.f. radiator, the factor becomes slightly less than 0.95. For most purposes, however, the figure of 0.95 may be taken as accurate. This assumes a radiator removed from surrounding objects, and with *no bends*.

Simple conversion into feet can be obtained by using the factor 1.56. To find the physical length of a half-wave 80-meter antenna, we multiply 80 times 1.56, and get 124.8 feet for the length of the radiator.

It is more common to use frequency than wavelength when indicating a specific spot in the radio spectrum. For this reason, the relationship between wavelength and frequency must be kept in mind. As the velocity of radio waves through space is constant at the speed of light, it will be seen that the more waves that pass a point per second (higher frequency), the closer together the peaks of those waves must be (shorter wavelength). Therefore, the higher the frequency the lower the wavelength.

A radio wave in space can be compared to a wave in water. The wave, in either case, has peaks and troughs. One peak and one trough constitute a *full wave,* or *one wavelength.*

Frequency describes the number of wave cycles or peaks passing a point per second. Wavelength describes the distance the wave travels through space during one cycle or oscillation of the antenna current; it is the distance in meters between adjacent peaks or adjacent troughs of a wave train.

As a radio wave travels 300,000,000 meters a second (speed of light), a frequency of 1 cycle per second corresponds to a wavelength of 300,000,000 meters. So, if the frequency is multiplied by a million, the wavelength must be divided by a million, in order to maintain their correct ratio.

A frequency of 1,000,000 cycles per second (1,000 kc.) equals a wavelength of 300 meters. Multiplying frequency by 10 and dividing wavelength by 10, we find: a frequency of 10,000 kc. equals a wavelength of 30 meters. Multiplying by 10 and dividing by 10 again, we get: a frequency of 100,000 kc. equals 3 meters wavelength. Therefore, to change wavelength to frequency (in kilocycles), simply divide 300,000 by the wavelength in meters.

$$F_{kc} = \frac{300{,}000}{\lambda}$$

$$\lambda = \frac{300{,}000}{F_{kc}}$$

Now that we have a simple conversion formula for converting wavelength to frequency and vice versa, we can combine it with our wavelength versus antenna length formula, and we have the following:

Wire length of a rectilinear half-wave radiator, in

$$\text{feet} = 1.56\lambda = \frac{467{,}400}{F_{kc}} = \frac{467.4}{F_{Mc}}$$

The slight discrepancy between the answers that will be obtained by the wavelength formula and by the frequency formula is due to the fact that the factor 1.56 is given only to two decimal places, this degree of accuracy being sufficient for ordinary purposes. Actually the factor is 1.558, but 1.56 is close enough, and simplifies calculations.

Harmonic Resonance. A wire in space resonates at more than one frequency. The *lowest* frequency at which it resonates is called its *fundamental* frequency, and at that frequency it is approximately a half wavelength long. A wire can have two, three, four, five, or more standing waves on it, and thus it resonates at approximately the integral harmonics of

its fundamental frequency. However, the higher harmonics are not exactly integral multiples of the lowest resonant frequency.

A harmonic operated antenna is somewhat longer than the corresponding integral number of dipoles, and for this reason, the dipole length formula cannot be used simply by multiplying by the corresponding harmonic. The intermediate half wave sections do not have "end effects." Also, the current distribution is disturbed by the fact that power can reach some of the half wave sections only by flowing through other sections, the latter then acting not only as radiators, but also as transmission lines. For the latter reason, the resonant length will be dependent to an extent upon the method of feed, as there will be less attenuation of the current along the antenna if it is fed at or near the center than if fed towards or at one end. Thus, the antenna would have to be somewhat longer if fed near one end than if fed near the center. The difference would be small, however, unless the antenna were at least 4 wavelengths long.

Under conditions of severe current attenuation, it is possible for some of the nodes, or loops, actually to be slightly greater than a physical half wavelength apart. It is obvious that with so many things affecting the length, the only method of resonating a harmonically operated antenna accurately is by cut and try, or by using a feed system in which both feed line and antenna are resonated at the station end as an integral unit.

A dipole or half-wave antenna is said to operate on its fundamental or first harmonic. A full wave antenna, 1 wavelength long, operates on its second harmonic. An antenna with five half-wavelengths on it would be operating on its fifth harmonic. Observe that the fifth harmonic antenna is $2\frac{1}{2}$ *wavelengths* long, *not* 5 wavelengths.

Antenna Impedance. In many ways, a half-wave antenna is like a tuned tank circuit. The main difference lies in the fact that the elements of inductance, capacity, and resistance are *lumped* in the tank circuit, and are *distributed* throughout the length of an antenna. The center of a half-wave radiator is effectively at ground potential as far as r.f. voltage is concerned, although the current is highest at that point. See Figure 2.

When the antenna is resonant, and it always should be for best results, the impedance at the center is a pure resistance, and is termed the radiation resistance. Radiation resistance is a fictitious term; it is that value of resistance (referred to the current loop) which would dissipate the same amount of power that is being radiated by the antenna.

The radiation resistance depends on the length of the antenna and its proximity to nearby objects which either absorb or reradiate power, such as the ground, other wires, etc.

Before going too far with the discussion of radiation resistance, an explanation of the Marconi (grounded quarter wave) antenna is in order. The Marconi antenna is a special type of Hertz antenna in which the earth acts as the "other half" of the dipole. In other words, the current flows into the earth instead of into a similar quarter-wave section. Thus, the current loop of a Marconi antenna is at the *base* rather than in the *center.* In either case, it is a quarter wavelength from the end (or ends).

A half-wave dipole far from ground and other reflecting objects has a radiation resistance at the center of about 73 ohms. Radiation resistance ordinarily is referred to a current loop. Otherwise, it has no particular significance, because it could be almost any value if the point on the antenna were not given.

A Marconi antenna is simply one-half of a dipole. For that reason, the radiation resistance is roughly half of 73 ohms.

Because the power throughout the antenna is the same, the *impedance* of the antenna at any point along its length merely expresses the ratio between voltage and current at that point. Thus, the lowest impedance occurs where the current is highest, namely, at the center of a dipole, or a quarter wave from the end of a Marconi. The impedance rises uniformly toward each end, where it is approximately 2400 ohms for a dipole remote from ground, and about twice as high for a vertical Marconi.

If a vertical half-wave antenna is set up so that its lower end is at the ground level, the effect of the ground reflection is to increase the radiation resistance to approximately 100 ohms. When a horizontal half-wave antenna is used, the radiation resistance (and, of course, the amount of energy radiated for a given antenna current) depends on the height of the antenna above ground, since the height determines the phase angle between the wave radiated directly in any direction and the wave which combines with it after reflection from the ground.

The radiation resistance of an antenna generally increases with length, although this increase varies up and down about a constantly increasing average. The peaks and dips are caused by the reactance of the antenna, when its length does not allow it to resonate at the operating frequency.

Antennas have a certain loss resistance as well as a radiation resistance. The loss re-

sistance defines the power lost in the antenna due to ohmic resistance of the wire, ground resistance (in the case of a Marconi), corona discharge, and insulator losses.

Resonance. Most antennas operate best when resonated to the frequency of operation. This does not apply to the terminated rhombic antenna, or to the *parasitic* elements of one popular type of close-spaced array, to be described later in the chapter. However, in practically every other case, it will be found that increased efficiency results when the entire antenna system is resonant, whether it be a simple dipole or an elaborate array. The radiation efficiency of a resonant wire is many times that of a wire which is not resonant.

If an antenna is slightly too long, it can be resonated by series insertion of a variable capacitor at a high current point. If it is slightly too short, it can be resonated easily by means of a variable inductance. These two methods are generally employed when part of the antenna is brought into the operating room.

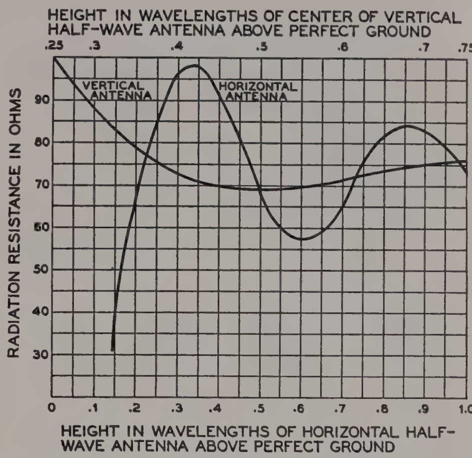
With an antenna array, or an antenna fed by means of a transmission line, it is more common to cut the elements to exact resonant length by "cut and try" procedure. Exact antenna resonance is more important when the antenna system has low radiation resistance; an antenna with low radiation resistance has higher Q (tunes sharper) than an antenna with high radiation resistance. The higher Q does not indicate greater efficiency; it simply indicates a sharper resonance curve.

CHARACTERISTICS AND CONSIDERATIONS

Radiation Resistance. Along a half-wave antenna, the *impedance* varies from a minimum at the center to a maximum at the ends. The impedance is that property which determines the antenna current at any point along the wire for the value of radio-frequency voltage at that point, assuming a given antenna power.

The curves of Figure 4 indicate the theoretical center-point radiation resistance of a half-wave antenna for various heights above perfect ground. These values are of importance in matching untuned radio-frequency feeders to the antenna, in order to obtain a good impedance match and an absence of standing waves on the feeders.

Above *average* ground, the actual radiation resistance of a dipole will vary from the exact value of Figure 4, since the latter assumes a hypothetical, perfect ground having no loss and perfect reflection. Fortunately, the curves for the radiation resistance over most types of



EFFECT OF HEIGHT ON THE RADIATION RESISTANCE OF A DIPOLE SUSPENDED ABOVE PERFECT GROUND.

earth will correspond rather closely with those of the chart, except that the radiation resistance for a horizontal dipole does not fall off as rapidly as is indicated for heights below an eighth wavelength. However, with the antenna so close to the ground and the soil in a strong field, much of the radiation resistance is actually represented by ground loss; this means that a good portion of the antenna power is being dissipated in the earth, which, unlike the hypothetical perfect ground, has resistance. In this case, an appreciable portion of the "radiation resistance" actually is loss resistance. The type of soil also has an effect upon the radiation *pattern,* especially in the vertical plane, as will be seen later.

When the radiation resistance of an antenna or array is very low, the current at a voltage node will be quite high for a given power. Likewise, the voltage at a current node will be very high. Even with a heavy conductor and excellent insulation, the losses due to the high voltage and current will be appreciable if the radiation resistance is sufficiently low.

Usually, it is not considered desirable to use an antenna or array with a radiation resistance of less than approximately 10 ohms unless there is sufficient directivity, compactness, or other advantage to offset the losses resulting from the low radiation resistance.

Ground Resistance. The radiation resistance of a Marconi antenna, especially, should be kept as high as possible. This will reduce the antenna current for a given power, thus minimizing loss resulting from the series resistance offered by the earth connection. The

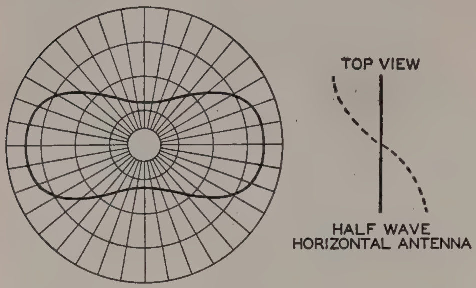


Figure 5.
RADIATION PATTERN OF A HALF-WAVE ANTENNA A HALF WAVE ABOVE PERFECT GROUND, FOR A FIXED VERTICAL ANGLE OF 30°.

radiation resistance can be kept high by making the Marconi radiator somewhat longer than a quarter wave, and shortening it by series capacity to an electrical quarter wave. This reduces the current flowing in the earth connection. It also should be removed from ground as much as possible (vertical being ideal). Methods of minimizing the resistance of the earth connection will be found in the discussion of the Marconi antenna.

Antenna Directivity

When choosing and orienting an antenna system, the radiation patterns of the various common types of antennas should be given careful consideration. The directional characteristics are of still greater importance when a directive antenna array is used.

There are two kinds of antenna directivity: vertical and horizontal. The latter is not generally desirable for amateur work except (1) for point-to-point work between stations regularly communicating with each other, (2) where several arrays are so placed as to cover most useful directions from a given location, and (3) when the beam may be directed by electrical or mechanical rotation.

Considerable horizontal directivity can be used to advantage for point-to-point work. Signals follow the great circle path, or are within 2 or 3 degrees of that path a good share of the time.

For general amateur work, however, *too much* horizontal directivity is ordinarily undesirable, inasmuch as it necessitates having the beam pointed exactly at the station being worked. Making the array rotatable overcomes this obstacle, but arrays having extremely high horizontal directivity are too cumbersome to be rotated, except perhaps above 56 Mc. The horizontal directivity of a horizontal dipole depends upon the vertical angle being considered. Directivity is greater for lower vertical angles. The polar diagram of Figure 5 shows a typical horizontal radiation pattern.

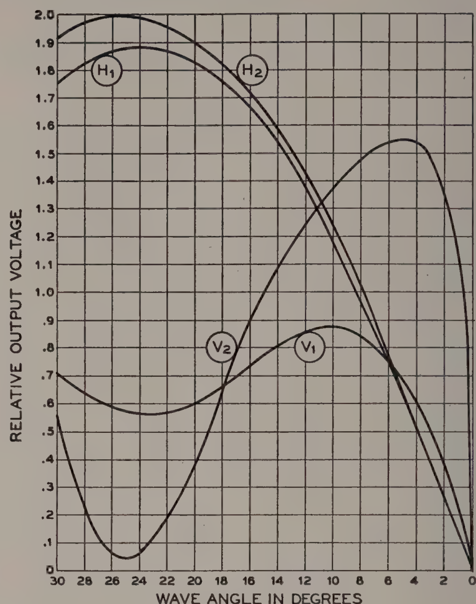


Figure 6.
VERTICAL - PLANE DIRECTIONAL CHARACTERISTICS OF HORIZONTAL AND VERTICAL DOUBLETS ELEVATED 0.6 WAVE-LENGTH AND ABOVE TWO TYPES OF GROUND.

H_1 represents a horizontal doublet over typical farmland, H_2 over salt water. V_1 is a vertical pattern of radiation from a vertical doublet over typical farmland, V_2 over salt water. A salt water ground is the closest approach to an extensive ideally perfect ground that will be met in actual practice.

On the 28- and 14-Mc. bands, and to an extent on the 7-Mc. band, the matter of vertical directivity is of as much importance as is horizontal directivity. Only the power leaving the antenna at certain vertical angles is instrumental in putting a signal into a distant receiving antenna; the rest may be considered as largely wasted. In other words, the important thing is the amount of power radiated in a desired direction *at the useful vertical angles,* rather than the actual shape of the directivity curves as read on the ground by a field strength meter, the latter giving only a pattern of the *ground wave.*

A nondirectional antenna, such as a vertical or horizontal dipole, will give excellent results with general coverage on 28 and 14 Mc. if the

vertical angle of radiation is favorable. The latter type is slightly directional broadside, especially on 28 Mc. where only very low angle radiation is useful, but nevertheless is considered as a "general coverage" type.

Effect of Average Ground on Antenna Radiation. Articles appearing in journals discussing antenna radiation often are based upon the perfect ground assumption, in order to cover the subject in the most simple manner. Yet, little has been said about the real situation which exists, the ground generally being anything but a perfect conductor. Consideration of the effect of a ground that is not perfect explains many things.

When the earth is less than a perfect conductor, it becomes a dielectric or, perhaps in an extreme case, a "leaky insulator."

The resulting change in the vertical pattern of a horizontal antenna is shown in Figure 6. The ground constants, in this case, are for flat farmland, which probably is similar to midwestern farmland. The ocean is the closest practical approach to a theoretically perfect ground. It will be noted that there is only a slight loss in power due to the imperfect ground as compared to the ocean horizontal.

The effect of the earth on the radiation pattern of a vertical dipole is much greater. Radiation from a half-wavelength vertical wire is severely reduced by deficiencies of the ground.

A very important factor in the advantages of horizontal or vertical dipoles, therefore, appears to be the condition of the ground.

The best angle of radiation varies with frequency, layer height, and many other factors. For instance, a lower optimum vertical angle is found to hold for high-frequency communication with South America from the U.S.A. than for Europe and the U.S.A.

FEEDING THE ANTENNA

Usually a high-frequency doublet or directional array is mounted as high and as much in the clear as possible, for obvious reasons. Power can then be fed to the antenna system via one of the various transmission lines discussed in the latter portion of this chapter.

However, it is sometimes justifiable to bring part of the radiating system directly to the transmitter, feeding the antenna without benefit of a transmission line. This is permissible when (1) there is insufficient room to erect a 75- or 160-meter horizontal dipole and feed line, (2) when a long wire is operated on one of the higher frequency bands on a harmonic. In either case, it is usually possible to get the main portion of the antenna in the clear because of its length. This means that the power lost by bringing the antenna directly to the transmitter is relatively small.

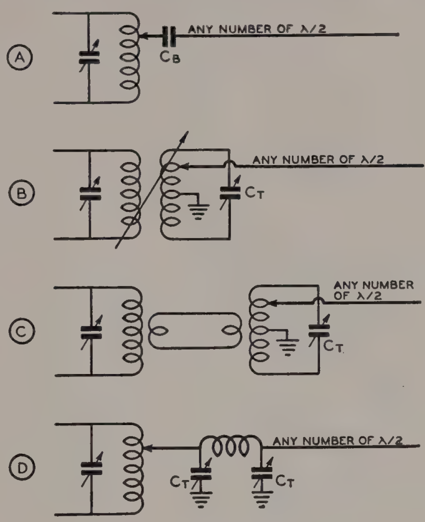


Figure 7.
FOUR METHODS OF END FEEDING AN ANTENNA.

The arrangement of "C" is to be recommended. The legality of arrangement "A" for amateur work is debatable if the blocking condenser is large. It is really a form of direct coupling, permitted by the regulations only when an untuned feed line is used.

Even so, it is not the best practice to bring the high-voltage end of an antenna into the operating room, especially for 'phone operation, because of the possibility of r.f. feedback from the strong antenna field. For this reason, one should dispense with a feed line in conjunction with a Hertz antenna only as a last resort.

End-Fed Antennas. The end-fed Fuchs (pronounced "Fooks") antenna has no form of transmission line to couple it to the transmitter, but brings the radiating portion of the antenna right down to the transmitter, where some form of coupling system is used to transfer energy to the antenna.

This antenna always is voltage-fed, and always consists of an *even* number of quarter-wavelengths. Figure 7 shows several common methods of feeding the Fuchs antenna or "end-fed Hertz." Arrangement "C" is to be recommended to minimize harmonics, as an end-fed antenna itself offers no discrimination against harmonics, either odd or even.

The Fuchs type of antenna has rather high losses unless at least three-quarters of the radiator can be placed outside the operating room and in the clear. As there is high r.f. voltage at the point where the antenna enters

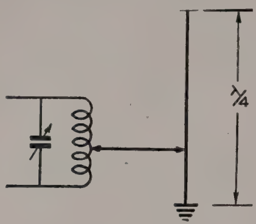


Figure 8.
THE SERIES-TUNED QUARTER-WAVE MARCONI, THE BASIC MARCONI ANTENNA SYSTEM.
The overall length to the earth connection, including lead in, is from 10% to 25% in excess of a quarter wavelength physically. The system is capacity-shortened and resonated by means of the series tuning condenser, with a maximum capacity of at least 3 μμfd. per meter of transmitter wavelength unless the antenna is considerably in excess of a quarter wave long.

the operating room, the insulation at that point should be several times as effective as the insulation commonly used with low-voltage feeder systems. This antenna can be operated on all of its higher harmonics with good efficiency, and can be operated at half frequency against ground as a quarter-wave Marconi.

As the frequency of an antenna is raised slightly when it is bent anywhere except at an exact voltage or current loop, a Fuchs antenna usually is a few per cent longer than a straight half-wave doublet for the same frequency, because, ordinarily, it is impracticable to bring a wire in to the transmitter without making several bends.

The Marconi Antenna

A grounded quarter-wave Marconi antenna is widely used on the 160-meter band, due to the fact that a half-wave antenna at that low frequency is around 260 feet long, which is out of the question for those confined to an ordinary city lot. It is also widely used by 1700–2500 kc. police services, and in u.h.f. mobile applications, where a compact radiator is required.

The Marconi type antenna allows the use of half of the length of wire used for a half-wave Hertz radiator. The ground acts as a mirror, in effect, and takes the place of the extra quarter wave that would be required to resonate the wire, were it not grounded.

The Marconi antenna generally is not as satisfactory for long distance communication as the Hertz type, and the radiation efficiency is never as great, due to the losses in the ground connection. However, it can be made almost as good a radiator on wavelengths longer than 120 meters, if *sufficient care is taken with the ground system.*

The fundamental Marconi antenna is shown in Figure 8, and all Marconi antennas differ from this only in the method of feeding energy. Antenna A in Figure 9 is the fundamental vertical type. Type B is the inverted-L type; type C is the T type, with the two halves of the top portion of the T effectively in parallel.

The Marconi antenna should be as *high as possible,* and too much attention cannot be paid to getting a low resistance ground connection.

Importance of Ground Connection. With a quarter-wave antenna and a ground, the antenna current generally is measured with a meter placed in the antenna circuit close to the ground connection. Now, if this current flows through a resistor, or if the ground itself presents some resistance, there definitely will be a power loss in the form of heat. Improving the ground connection, therefore, provides a definite means of reducing this loss of antenna power, and thus increasing the radiated power.

The best possible ground consists of as many wires as possible, each at least a quarter wave long, buried just below the surface of the earth, and extending out from a common point in the form of radials. Copper wire of any size larger than no. 16 is satisfactory, though the larger sizes will take longer to disintegrate. In fact, the radials need not even be buried; they may be supported just above the earth, and insulated from it. This arrangement is called a *counterpoise,* and operates by virtue of its high capacity to ground.

Unless a large number of radials is used, fairly close to the ground, the counterpoise will act more like the bottom half of a half-wave Hertz than like a ground system. However, the efficiency with a counterpoise will be quite good, regardless. It is when the radials are buried, or laid on the ground, that a large number should be used for best efficiency. Broadcast stations use as many as 120 radials of from 0.3 to 0.5 wavelength long.

A large number of radials not only provides a low resistance earth connection, but also, if long enough, produces the effect of locating the radiator over highly conducting earth. The importance of the latter with regard to vertical antennas is illustrated in Figure 6.

When it is impossible to extend buried radials in all directions from the ground connection for an inverted-L type Marconi, it is of importance that a few wires be buried directly below the flat top, and spaced at least 10 feet from one another.

If the antenna should be physically shorter than a quarter wavelength, antenna current would be higher, due to lower radiation resistance. Consequently, the power lost in resistive soil would be greater. The importance of a good ground with short, inductive-loaded Marconi radiators is, therefore, quite obvious. With a good ground system, even very short (one-eighth wavelength) antennas can be expected to give upwards from 90 per cent of the efficiency of a quarter-wave antenna used with the same ground system. This is especially true when the short radiator is *top loaded* with a high Q (low loss) coil. In this type radiator, the loading coil is placed near the top of the radiator, rather than at the bottom.

Water Pipe Grounds. Water pipe, because of its comparatively large surface and cross section, has as low an r.f. resistance as copper wire. If it is possible to attach to a junction of several water pipes (where they branch in several directions and run for some distance under ground), a satisfactory ground connection will be obtained. If one of the pipes attaches to a lawn or garden sprinkler system in the immediate vicinity of the antenna, and runs hither and thither to several neighboring faucets within a radius of a hundred yards, the effectiveness of the system will approach that of buried copper radials.

The main objection to water pipe grounds is the possibility of high resistance joints in the pipe, due to the "dope" put on the coupling threads. By attaching the ground wire to a junction with three or more legs, the possibility of requiring the main portion of the r.f. current to flow through a high resistance connection is greatly reduced.

The presence of water in the pipe adds but little to the conductivity; therefore it does not relieve the problem of high resistance joints. Bonding the joints is the best insurance, but this is, of course, impracticable where the pipe is buried. Bonding together with copper wire the various water faucets above the surface of the ground will improve the effectiveness of a water pipe ground system hampered by high resistance pipe couplings.

Marconi Dimensions. A Marconi antenna is exactly an odd number of electrical quarter waves long (usually only one quarter wave in length), and is always resonated to the operating frequency. The correct loading of the final amplifier is accomplished by varying the coupling, *rather than by detuning the antenna from resonance.*

Physically, a quarter-wave Marconi may be made anything from one-eighth to three-eighths wavelength overall, meaning the total length of the antenna wire and ground lead

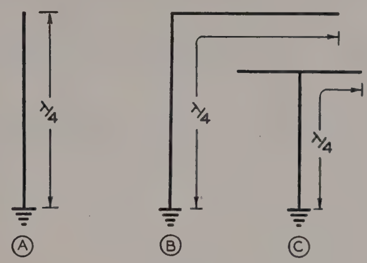


Figure 9.

THREE COMMON VARIATIONS OF THE MARCONI ANTENNA.

The bottom half of the radiator does most of the radiating, regardless of which type is used, because the current is greatest through that portion of the antenna.

from the end of the antenna to the point where the ground lead attaches to the junction of the radials or counterpoise wires, or the water pipe enters the ground. The longer the antenna is made physically, the lower will be the current flowing in the ground connection, and the greater will be the overall radiation efficiency. However, when the antenna length exceeds three-eighths wavelength, the antenna becomes difficult to resonate by means of a series condenser, and it begins to take shape as an end-fed Hertz, requiring a different method of feed than that illustrated in Figure 8 for current feed of a Marconi.

A radiator physically shorter than a quarter wavelength can be lengthened electrically by means of a series loading coil, and used as a quarter-wave Marconi. However, if the wire is made shorter than approximately one-eighth wavelength, the radiation resistance will be so low that high efficiency cannot be obtained, even with a very good ground.

Loading Coils. To resonate inductively an inductive-loaded Marconi, the inductance would have to be in the form of a variometer in order to permit continuous variation of the inductance. The more common practice is to use a tapped loading coil and a series tuning condenser. The loading coil should preferably be placed a short distance from the *top or far end* of the radiator; this reduces the current flowing in the ground connection by raising the radiation resistance, resulting in better radiation efficiency. More than the required amount of inductance for resonance is clipped in series with the antenna, and the system is then resonated by means of the series variable condenser, the same as though the radiator were actually too long physically.

To estimate whether a loading coil will probably be required, it is necessary only to note

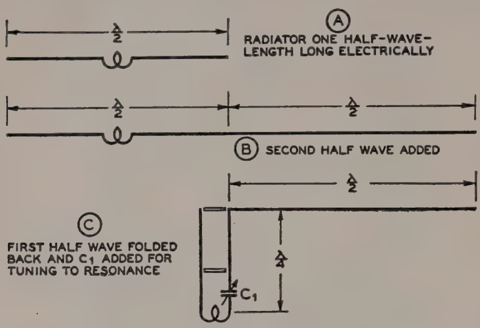


Figure 10.
THE EVOLUTION OF A ZEPP ANTENNA.

if the length of the antenna wire and ground lead is over a quarter wavelength; if so, no loading coil should be required, provided the series tuning condenser has a high maximum capacity.

Amateurs primarily interested in the higher frequency bands, but who like to work 160 meters occasionally, can usually manage to resonate one of their antennas as a Marconi by working the whole system, feeders and all, against a water pipe ground, and resorting to a loading coil if necessary. A high-frequency zepp, doublet, or single-wire-fed antenna will make quite a good 160-meter Marconi if high and in the clear, with a rather long feed line to act as a radiator on 160 meters. Where two-wire feeders are used, the feeders should be tied together for Marconi operation.

TRANSMISSION LINES

For many reasons, it is desirable to place a radiator as high and as much in the clear as possible, utilizing some form of nonradiating transmission line to carry energy with as little loss as possible from the transmitter to the radiating antenna.

There are many different kinds of transmission lines, and, generally speaking, practically any type of transmission line or feeder system can be used with any type of antenna; however, certain types are often better adapted than others for use with a certain antenna.

Transmission lines are of two general types: resonant and nonresonant. Strictly speaking, the term *transmission line* should really only be applied to a *nonresonant line*. Strictly speaking, a *resonant* line should be termed a *feeder* system, such as zepp feeders, etc. However, *transmission line* has come to refer to either type of line, tuned or untuned.

The principal types of nonresonant transmission lines include the single-wire-feed, the two-wire open and the twisted-pair matched impedance, the coaxial (concentric) feed line, and the multi-wave matched-impedance open line.

Voltage Feed and Current Feed. The half-wave Hertz antenna has high voltage and low current at each end, and it has low voltage and high current at its center. As any ungrounded resonant antenna consists merely of one or more half-wave antennas placed end to end, it will be seen that there will be a point of high r.f. voltage approximately every half wave of length measured from either end of the antenna. Also, there will be a point of high r.f. current half-way between any two adjacent high voltage points.

A voltage-fed antenna is any antenna which is excited at one of these high voltage points, or, in other words, a point of high impedance. Likewise, a current-fed antenna is one excited at a point along the antenna where the current is high and the voltage low, which corresponds to a point of low impedance.

The Zepp Antenna

The zepp antenna system is easy to tune up, and can be used on several bands by merely retuning the feeders. The overall efficiency of the zepp antenna system is probably not quite as high for long feeder lengths as for some of the antenna systems which employ nonresonant transmission lines, but where space is limited and where operation on more than one band is desired, the zepp has some decided advantages.

Zepp feeders really consist of an additional length of antenna which is folded back on itself, so that the radiation from the two halves cancels. In Figure 10A is shown a simple Hertz antenna, fed at the center by means of a pickup coil. Figure 10B shows another half-wave radiator tied directly on one end of the radiator shown in Figure 10A. Figure 10C is exactly the same thing, except that the first half-wave radiator, in which is located the coupling coil, has been folded back on itself. In this particular case, each half of the folded part of the antenna is exactly a quarter-wave long electrically.

Addition of the coupling coil naturally will electrically lengthen the antenna; thus, in order to bring this portion of the antenna back to resonance, we must electrically shorten it by means of the series tuning condenser, C_1. The two wires in the folded portion of the antenna system do not have to be exactly a quarter wave long physically, although the total *electrical* length of the folded portion must be equal to one-half wavelength electrically.

When the total electrical length of the two feeder wires, plus the coupling coil, is slightly greater than any odd multiple of one-quarter wave, then series condensers must be used to shorten the electrical length of the feeders sufficiently to establish resonance. If, on the other hand, the electrical length of the feeders and the coupling coil is slightly less than any odd multiple of one-quarter wave, then parallel tuning (wherein a condenser is shunted across the coupling coil) must be used in order to increase the electrical length of the whole feeder system to a multiple of one-quarter wavelength.

As the radiating portion of the zepp antenna system must always be some multiple of a half wave long, there is always high voltage present at the point where the live zepp feeder attaches to the end of the radiating portion of the antenna. Thus, this type of zepp antenna system is *voltage fed.*

The idea that it takes two condensers to balance the current in the feeders, one condenser in each feeder, is a common misconception regarding the zepp type end-fed antenna. Balancing the feeders with tuning condensers for equal currents is useless, anyhow, inasmuch as the feeders on an end-fed zepp can never be balanced for *both* current *and* phase because of the tendency for the end of the "dead" feeder to have more voltage on it than the one attached to the radiator.

Flat Top Length. The correct physical length for the flat top (radiating portion) of a zepp is *not* 0.95 of a half wavelength. Instead, it is so close to a half wavelength that it may be taken as that figure. Thus, while a 7300-kc. doublet is 64 feet long, the flat top of a 7300-kc. *zepp* should be 67 feet 3 inches. The reason for this is readily apparent when it is remembered that the 5 per cent difference between a resonant doublet and a physical half wavelength is principally due to "end effects," 2½ per cent at each end of the radiator.

Obviously there is no end effect at the end of a radiator to which zepp feeders are attached. Hence, we lengthen the radiator 2½ per cent. Now we must take into consideration that the end of the "dead" (unattached) feed wire has end effects, and that the other feeder does not. We want the two voltage loops to come at the same point on the feed line in order to obtain the best possible balance so as to minimize radiation. So we make the dead feed wire 2½ per cent of a half wavelength shorter than the other. This can be done quite easily, merely by lengthening the flat top another 2½ per cent. Thus, as the reader can readily see, the flat top is 5 per cent longer than if it were fed in the center.

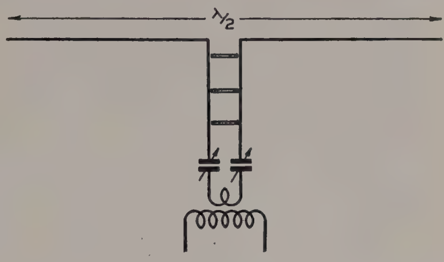


Figure 11.

THE TUNED DOUBLET USES AN OPEN-WIRE FEED SYSTEM.

The flat top need not be exactly an electrical half wave in length so long as the whole system, both flat top and feeders, is resonant as a unit. Only one tuning condenser need be used if desired. Certain feeder lengths will require that the condenser be placed across the coil rather than in series with it.

The Tuned Doublet

A current-fed doublet with spaced feeders, sometimes erroneously called a center-fed zepp, is an inherently balanced system (if the two legs of the radiator are exactly equal electrically), and there will be no radiation from the feeders regardless of what frequency the system is operated on. A series condenser may be put in *one* feeder (if right at the coupling coil) without affecting the balance of the system. The system can successfully be operated on almost any frequency, if the system as a whole can be resonated to the operating frequency. This is usually possible with a tapped coil and a tuning condenser that can optionally be placed either across the antenna coil or in series with it.

This type of antenna system is shown in Figure 11. It is a current-fed system on the lowest frequency for which it will operate, but becomes a voltage-fed system on all its even harmonics.

The antenna has a different radiation pattern when operated on harmonics, as would be expected. The arrangement used on the second harmonic is better known as the Franklin colinear array, and is described later in this chapter. The pattern is similar to a half-wave doublet, except that it is sharper in the broadside direction. On harmonics there will be multiple lobes, but the minor lobes are quite small as compared to the major lobe.

Tuned Feeder Considerations

If a transmission line is terminated in its *characteristic surge impedance,* there will be no reflection at the end of the line, and the

current and voltage distribution will be uniform along the line. If the end of the line is either open-circuited or short-circuited, the reflection at the end of the line will be 100 per cent, and *standing waves* of very great amplitude will appear on the line. There will still be practically no radiation from the line, but voltage nodes will be found along the line, spaced a half wavelength. Likewise, voltage loops will be found every half wavelength, the voltage loops corresponding to current nodes.

If the line is terminated in some value of resistance other than the characteristic surge impedance, there will be some reflection, the amount being determined by the amount of mismatch. With reflection, there will be standing waves (excursions of current and voltage) along the line, though not to the same extent as with an open-circuited or short-circuited line. The current and voltage loops will occur at the same *points* along the line as with the open- or short-circuited line, and as the terminating impedance is made to approach the characteristic impedance of the line, the current and voltage along the line will become more uniform. The foregoing assumes, of course, a purely resistive (nonreactive) load.

A well built 500- to 600-ohm transmission line may be used as a resonant feeder for lengths up to several hundred feet with very low loss, so long as the amplitude of the standing waves (ratio of maximum to minimum voltage along the line) is *not too great*. The amplitude, in turn, depends upon the mismatch at the line termination. A line of no. 12 wire, spaced 6 inches with good ceramic or Lucite spreaders, has a surge impedance of approximately 600 ohms, and makes an excellent tuned feeder for feeding anything between 60 and 6000 ohms (at frequencies below 30 Mc.). If used to feed a load of higher or lower impedance than this, the standing waves become great enough in amplitude that some loss will occur unless the feeder is kept short. At frequencies above 30 Mc., the spacing becomes an appreciable fraction of a wavelength, and radiation from the line no longer is negligible.

If a transmission line is not perfectly matched, it should be made *resonant*, even though the amplitude of the standing waves (voltage variation) is not particularly great. This prevents reactance from being coupled into the final amplifier. A feed system having moderate standing waves may be made to present a nonreactive load to the amplifier either by tuning or by pruning the feeders to approximate resonance.

Usually it is preferable with tuned feeders to have a current loop (voltage minimum) at the transmitter end of the line. This means that when voltage-feeding an antenna, the tuned feeders should be made an odd number of quarter wavelengths long, and when current-feeding an antenna, the feeders should be made an even number of quarter wavelengths long. Actually, the feeders are made about 10 per cent of a quarter wave longer than the calculated value (the value given in the tables) when they are to be series tuned to resonance by means of a condenser, instead of being trimmed and pruned to resonance.

When tuned feeders are used to feed an antenna on more than one band, it is necessary to compromise and make provision for both series and parallel tuning, inasmuch as it is impossible to cut a feeder to a length that will be optimum for several bands. If a voltage loop appears at the transmitter end of the line on certain bands, parallel tuning of the feeders will be required in order to get a transfer of energy. It is impossible to transfer energy by inductive coupling unless current is flowing. This is effected at a voltage loop by the presence of the resonant tank circuit formed by parallel tuning of the antenna coil.

Methods of coupling to a transmitter are discussed later in the chapter.

Untuned Transmission Lines

A nonresonant or untuned line is a line with negligible standing waves. Physically, the line itself should be *identical throughout its length*. There will be a smooth distribution of voltage and current throughout its length, both tapering off very slightly towards the antenna end of the line as a result of line losses. The attenuation (loss) in certain types of untuned lines can be kept very low for line lengths up to several thousand feet. In other types, particularly where the dielectric is not air (such as in the twisted-pair line), the losses may become excessive at the higher frequencies, unless the line is relatively short.

The termination at the antenna end is the only critical characteristic about the untuned line. It is the reflection from the antenna end which starts waves moving back toward the transmitter end. When waves moving in both directions along a conductor meet, standing waves are set up.

All transmission lines have distributed inductance, capacity, and resistance. Neglecting the resistance, as it is of minor importance in short lines, it is found that the *inductance and capacity per unit length* determine the characteristic or surge impedance of the line. Thus, the surge impedance depends upon the nature and spacing of the conductors, and the dielectric separating them.

When any transmission line is terminated in an impedance equal to its surge impedance,

reflection of energy does not occur, and no standing waves are present. When the load termination is exactly the same as the line impedance, it simply means that the load takes energy from the line just as fast as the line delivers it, no slower and no faster.

Thus, for proper operation of an untuned line (with standing waves eliminated), some form of impedance-matching arrangement must be used between the transmission line and the antenna, so that the radiation resistance of the antenna is reflected back into the line as a nonreactive impedance equal to the line impedance. It is important that the *radiator itself be cut to exact resonance;* otherwise, it will not present a pure resistive load to the nonresonant line.

An untuned feeder system may consist of one, two, four, or even more parallel wires. Increased constructional difficulties of the multi-wire type of line, where three or more parallel wires are used, and there is danger of appreciable feeder radiation from an improperly adjusted single-wire feeder, make the more familiar two-wire type of line the most satisfactory for general use.

Semi-Resonant Open Lines. As has been previously stated under *Tuned Feeder Considerations,* a well built *open-wire* line has low losses, even when standing waves with a ratio of as high as 10/1 are present. (The standing wave ratio will be found to approximate the ratio of mismatch at the feeder termination.) Of much greater importance is to make sure the line is *balanced,* which means that the antenna system must be electrically symmetrical, or allowance made for the asymmetry. If the currents in the two feed wires are not equal in amplitude and exactly opposite in phase, there will be radiation from the line (or pickup by the line if used for receiving), regardless of the amplitude of standing waves.

Because moderate standing waves can be tolerated on open-wire lines without loss, a standing wave ratio of 2/1 or 3/1 is considered acceptable with this type of line, *even when used in an "untuned" system.* Strictly speaking, a line is untuned, or nonresonant, only when the line is perfectly "flat," with a standing wave ratio of 1 (no standing waves). However, some mismatch can be tolerated with open-wire untuned lines, so long as the reactance is not objectionable, or else is eliminated by cutting the line to approximately a resonant length.

Thus, we have a line that is a cross between a tuned and an untuned line. Most of the "untuned" open-wire lines used by amateurs fall in this class, because there is usually more or less of a mismatch at the line termination. Therefore, *open-wire* lines with a standing wave ratio of *less than* 3/1 may be classed as *nonresonant,* or untuned, lines, as standing waves will not affect the operation of an untuned line unless greater than this in magnitude.

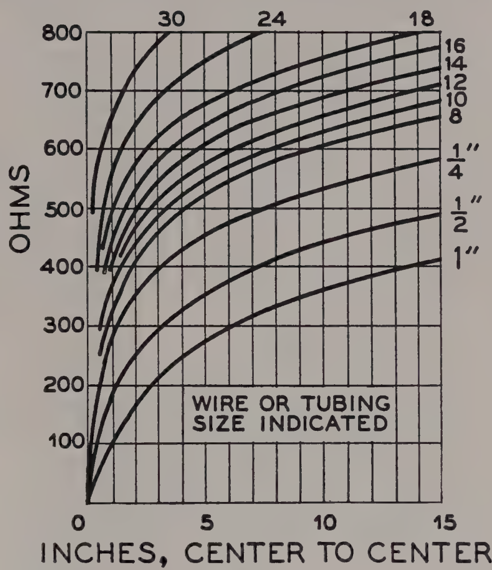


Figure 12.

CONDUCTOR SIZE AND SPACING VERSUS SURGE IMPEDANCE FOR TWO-WIRE OPEN LINE OR MATCHING TRANSFORMER.

The foregoing applies only to *open-wire lines.* The losses in other type lines, especially those having rubber dielectric, go up rapidly with the standing wave ratio, such lines being designed for perfectly "flat" operation. Also, the maximum power handling capability of lines is greatly reduced when standing waves are present, even though of only 2/1 or 3/1 magnitude. The power handling capability of an open line will still be very high, but other lines do not have such a high capability to begin with, and if being worked at full rated power may be punctured by the presence of moderate standing waves. From this we can see that every attempt should be made to eliminate all traces of standing waves on a low impedance, close-spaced line, especially when the power is high enough that there is danger of arc-over at voltage loops, or when the frequency is high enough that the losses are already so great that increased losses will be a serious item.

Construction of Two-Wire Open Lines. A two-wire transmission system is easy to construct. Its surge impedance can be calcu-

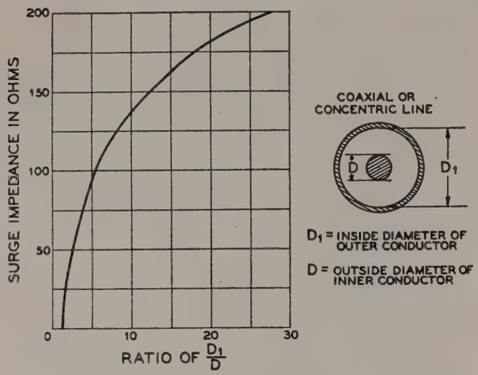


Figure 13.
CURVE FOR DETERMINATION OF SURGE IMPEDANCE OF ANY COAXIAL LINE HAVING AIR DIELECTRIC.
Presence of spacing insulators will lower the impedance somewhat below the calculated value as derived from this chart.

lated quite easily, and when properly adjusted and balanced to ground, undesirable feeder radiation is minimized; the current flow in the adjacent wires is in opposite directions, and the magnetic fields of the two wires are in opposition to each other. When a two-wire line is terminated with the equivalent of a pure resistance equal to the surge impedance of the line, the line becomes a nonresonant line.

It can be shown mathematically that the true surge impedance of any two-wire parallel line system is approximately equal to

$$Z_s = 276 \log_{10} \frac{2S}{d}$$

Where:

S is the exact distance between wire centers in some convenient unit of measurement, and

d is the diameter of the wire measured in the same units as the wire spacing, S.

Since $\frac{2S}{d}$ expresses a ratio only, the units of measurement may be centimeters, millimeters, or inches. This makes no difference in the answer, so long as the substituted values for S and d are in the *same units*.

The equation is accurate *so long as the wire spacing is relatively large as compared to the wire diameter*.

Surge impedance values of less than 200 ohms are seldom used in the open-type two-wire line, and, even at this comparatively high

COMPARATIVE R. F. FEEDER LOSSES

FRE-QUENCY	DB LOSS PER 100 FT.	TYPE OF LINE
7 Mc. 14 Mc. 30 Mc.	0.9 1.5 3	70-ohm impedance, rubber insulated twisted-pair with outer covering of braid.
7 Mc. 30 Mc. 60 Mc.	0.6 0.9 1.3	⅜" concentric pipe feeder with inner wire on bead spacers. Impedance, 70 ohms.
7 Mc. 30 Mc. 60 Mc.	0.05 0.12 0.18	Open 2-wire line no. 10 wire, 2 in. spacing.
7 Mc. 14 Mc. 30 Mc.	3 4.5 8	Twisted no. 14 solid weatherproof wire, weathered for six months (telephone wire).
7 Mc. 14 Mc. 30 Mc.	1.5 2.5 4.2	Heavy, flexible coaxial cable, rubber insulation, metal braid outer conductor.

value of Z_s, the wire spacing S is uncomfortably close, being only 5.3 times the wire diameter d.

Figure 12 gives in graphical form the surge impedance of any practicable two-wire line. The chart is self-explanatory, and sufficiently accurate for practical purposes.

Twisted-Pair Untuned Lines. Low-loss, low-impedance transmission cable, marketed by several manufacturers under the trade name of "EO-1 cable," allows a very flexible transmission line system to be used to convey energy to the antenna from the transmitter. The low-loss construction is largely due to the use of untinned solid conductors, low-loss insulation, plus a good grade of weatherproof covering.

Twisted no. 12 or no. 14 outside house wire may be used on 160 and 80 meters if the length is not over 50 or 75 feet. On higher frequencies, however, the losses with such "homemade" twisted line will be excessive.

A twisted-pair line should always be used as an untuned line, as standing waves on the line will produce excessive losses, and can easily break down the line insulation at the voltage loops.

For turning sharp corners and running close to large bodies of metal, the twisted pair is almost as good at the lower frequencies as the coaxial line.

Above 14 Mc., however, the rubber insulation causes appreciable dielectric loss even

with the best EO-1 cable, and the twisted-pair type of low-impedance line should not be used except where the length is short, or where more efficient lines might not be suitable from a mechanical standpoint, as in certain types of rotary arrays.

The low surge impedance of the twisted-pair transmission line is due not only to the close spacing of the conductors, but to the rubber insulation separating them. The latter has a dielectric constant considerably higher than that of air. This not only lowers the surge impedance but also results in slower propagation of a wave along the conductors. As a result, the voltage loops occur closer together on the line when standing waves are present than for an open-wire line working at the same frequency.

Coaxial Line. Several types of coaxial cable have come into wide use for feeding power to an antenna system. A cross-sectional end view of a coaxial cable (sometimes called concentric cable or line) is shown in Figure 13.

As in the parallel-wire line, the power lost in a properly terminated coaxial line is the sum of the effective resistance losses along the length of the cable and the dielectric losses between the two conductors. In a well designed line using air or nitrogen as the dielectric, both are negligible, the actual measured loss in a good line being less than 0.5 db per 1000 feet at 1 megacycle.

Of the two losses, the effective resistance loss is the greater; since it is largely due to the skin effect, the line loss (all other conditions the same) will increase directly as the square root of the frequency.

Figure 13 shows that, instead of having two conductors running side by side, one of the conductors is placed *inside* of the other. Since the outside conductor completely shields the inner one, no radiation takes place. The conductors may both be tubes, one within the other, or the line may consist of a solid wire within a tube.

In one type of cable (solid or semi-flexible low-loss type), the inner conductor is supported at regular intervals from the outside tube by a circular insulator of either pyrex, polystyrene, or some non-hygroscopic ceramic material with low high-frequency losses. The insulators are slipped over the inner conductor, and held in place either by some system of small clamps, or by crimping the wire immediately in front of and behind each insulator.

Moisture must be kept out of the tube if best results are to be secured. For this reason, it is necessary to solder or otherwise tightly join the line sections together so that no leak occurs. This prevents water from seeping into the line in outdoor installations.

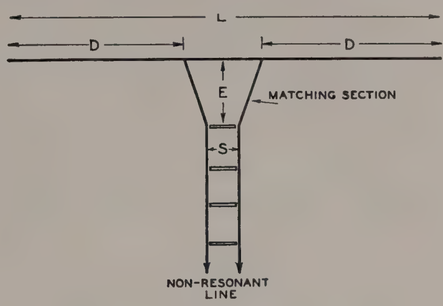


Figure 14.
THE DELTA-MATCHED ANTENNA SYSTEM.
This system is sometimes called a "Y matched" doublet. For dimensions refer to formula in text.

To avoid condensation of moisture on the inside walls of the line, it is general commercial practice to fill the line with dry nitrogen gas at a pressure of approximately 35 pounds per square inch.

Filling a line with dry nitrogen gas also greatly increases its power capacity, a power capability rating of 3 to 1 being quite common for the nitrogen-filled line as compared to a line operating under normal atmospheric pressures.

Nearby metallic objects cause no loss, and the cable may be run up air ducts, wire conduit, or elevator shafts. Insulation troubles can be forgotten. The coaxial cable may be either buried in the ground or suspended above ground.

Highly flexible coaxial cable having continuous rubber dielectric for maintenance of spacing, and an outer conductor of shield braid of the type used for ordinary shielded wire, has become quite popular for certain applications where cost is an item, or the line is subjected to continual flexing. Because of the rubber dielectric, the losses are about the same as for EO-1 cable on the higher frequencies, while on the lower frequencies (below 4000 kc.) the losses are nearly as low as for the air-dielectric type of coaxial line.

The chief advantage of rubber dielectric coaxial cable over EO-1 cable is its availability in lower values of surge impedance, making it possible to feed certain types of low radiation resistance arrays without need for an impedance matching device. Twisted-pair cable it not commonly available with a surge impedance of less than 70 ohms, while rubber dielectric coaxial cable is available with a surge impedance of as low as 28 ohms.

Coaxial cable, like twisted-pair cable, is most commonly used without a matching system.

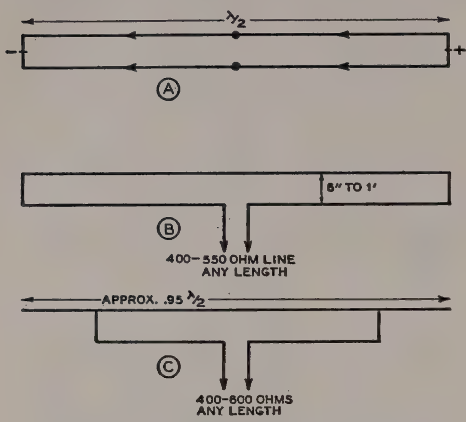


Figure 15.

The two wire doublet, shown at A, can be fed in one leg as shown at B in order to provide high terminal radiation resistance. Arrows indicate current flow on one half the cycle, dots indicate position of current loops. The modified version due to Kraus, shown at C, provides a more accurate match for single frequency operation when the line impedance is over 300 ohms. The latter arrangement is referred to as the "T matched" antenna.

Cable is chosen to have a surge impedance that approximates the terminal radiation resistance of the antenna (point at which the line is connected).

While coaxial cable is best suited to use with Marconi antennas, because the outside conductor is ordinarily grounded, it can be used successfully to feed a balanced dipole. This is permissible because the impedance is low, and therefore no great unbalance results from such operation. The outer conductor of the coaxial cable connects to one half the dipole, and the inner conductor connects to the other half. In this case, the outer conductor is often left ungrounded.

Matching Nonresonant Lines to the Antenna

From the standpoint of economy and efficiency, the most practical untuned line is an open line having a surge impedance of from 440 to 600 ohms. Unfortunately, it is seldom that the antenna system being fed has an impedance of similar value either at a current loop or at a voltage loop. It is sometimes necessary, with current-fed antennas, to match the line to an impedance as low as 8 or 10 ohms, while with voltage-fed antenna systems and arrays, it is occasionally necessary to match the line to an impedance of many thousands of ohms. There are many ways of accomplishing this, the more common and most satisfactory methods being discussed here.

Delta-Matched Antenna System. The delta type matched-impedance antenna system is shown in Figure 14. The impedance of the transmission line is transformed gradually into a higher value by the fanned-out Y portion of the feeders, and the Y portion is tapped on the antenna at points where the antenna impedance is a compromise between the impedance at the ends of the Y and the impedance of the unfanned portion of the line.

The constants of the system are rather critical, and the antenna must resonate at the operating frequency in order to minimize standing waves on the line. Some slight readjustment of the taps on the antenna is desirable, if appreciable standing waves persist in appearing on the line. It is almost impossible to get the standing wave ratio below 2/1 with this system, and as standing waves of this order are not objectionable on an open line if it is cut to such a length that it is non-reactive, this ratio is considered as indicating the best match that can be expected with a "Y" or delta-matched doublet.

The constants are determined by the following formulas:

$$L_{feet} = \frac{467.4}{F \text{ megacycles}}$$

$$D_{feet} = \frac{175}{F \text{ megacycles}}$$

$$E_{feet} = \frac{147.6}{F \text{ megacycles}}$$

where L is antenna length; D is the distance *in from each end* at which the Y taps on; E is the height of the Y section.

As these constants are correct only for a 600-ohm transmission line, the spacing S of the line must be approximately 75 times the diameter of the wire used in the transmission line. For no. 14 B & S wire, the spacing will be slightly less than 5 inches. For no. 12 B & S, the spacing should be 6 inches. This system should never be used on either its even or odd harmonics, as entirely different constants are required when more than a single half wavelength appears on the radiating portion of the system.

The Multi-Wire Doublet. When a doublet consists of two or more wires instead of the more usual single wire, the radiation resistance (impedance at the current loop) is raised slightly. This is due to the fact that each wire tends to induce an opposing current in the

opposite wire, but cannot because the two wires are tied together at either end. See "A," Figure 15.

If we split just one wire of such an antenna, as at "B" in Figure 15, and feed the antenna at this point, we find that the *terminal* radiation resistance is much higher than the theoretical 72 ohms of a conventional doublet. The terminal radiation resistance is the impedance into which the feed system works. Because each wire of the two-wire doublet carries half the total current, and the feed line serves only one wire, the terminal radiation resistance is four times the radiation resistance of the antenna taken as a whole, which already is slightly higher than that of a regular doublet.

The terminal radiation resistance of a two-wire doublet such as that of "B" when well removed from earth is about 300 ohms. This permits use of an ordinary 500 to 600 ohm open line to feed the antenna directly, without need for a matching system. When used with a 500-ohm line (no. 12 spaced 4 inches) the standing waves will be quite low (approximately 2/1 ratio) over a range in frequency of several per cent either side of resonance. The broad tuning characteristic is a result of the high radiation resistance.

The spacing of the two wires is not at all critical, and need not be exactly uniform so long as the system is symmetrical. The overall length of the "loop" is one wavelength. The lower element is split in the exact center for attachment to the feed line.

A useful variation of the two-wire doublet is the "T matched" doublet, which is a cross between a delta matched doublet and a two-wire doublet. It may be considered as a delta matched system without the usual fanning of the feed line at point of attachment. This antenna is illustrated at Figure 15-C. When both radiator length and point of feeder attachment are exactly right (as determined experimentally), the standing waves will be so low as to be almost negligible.

Single Wire Fed Antenna. If one wire is removed from the delta matched impedance antenna of Figure 14 and the remaining feeder is moved along the doublet to the point giving the lowest standing wave ratio on the single feed wire, the system will still work satisfactorily. However, there will be an appreciable amount of radiation from the feeder, even with the best possible match, and for this reason a single wire feeder is never used to feed directive antenna arrays, and is used primarily for portable and emergency work.

A single-wire feed line has a characteristic surge impedance of from 500 to 600 ohms, depending upon the diameter of the feeder wire. This type feeder makes use of the earth as a return circuit through the earth's capacity effect to the antenna and feeder. The actual earth connection to the transmitter may have a relatively high resistance without causing appreciable loss of r.f energy. It may even be represented by the capacity of the transmitter and house wiring to earth.

The feeder is normally attached to the radiator about $\frac{1}{6}$ or $\frac{1}{7}$ of a half wavelength from the center.

The single wire fed antenna not only works well on its fundamental, but is a good radiator on its various harmonics. For this reason, this type antenna system should not be used on the low- and medium-frequency bands, unless a harmonic suppressing antenna coupler is used to prevent radiation of harmonics.

A single wire feeder also can be used to feed a quarter wave vertical Marconi radiator. The best point of attachment for the feeder should be determined by cut and try. Normally it will be about $\frac{1}{3}$ of the way up the radiator.

Matching Stubs

By hanging a resonant length of Lecher wire line (called a matching stub) from either a voltage or current loop and attaching 600-ohm nonresonant feeders to the resonant stub at a suitable voltage (impedance) point, standing waves on the line may be virtually eliminated. The stub is made to serve as an autotransformer. Thus, by putting up a half-wave zepp with quarter-wave feeders at a distance from the transmitter, and attaching a 600-ohm line from the transmitter to the zepp feeders at a suitable point, we have a stub-matched antenna. The example cited here is commonly called a J antenna, especially when both radiator and stub are vertical. Many variations from this example are possible; stubs are particularly adapted to matching an open line to certain directional arrays, as will be described later in this chapter.

Voltage Feed. When the stub attaches to the antenna at a voltage loop, the stub should be a quarter wavelength long electrically, and be shorted at the bottom end. The stub can be resonated by sliding the shorting bar up and down before the nonresonant feeders are attached to the stub, the antenna being shock-excited from a separate radiator during the process. Slight errors in the length of the radiator can be compensated for by adjustment of the stub if both sides of the stub are connected to the radiator in a symmetrical manner. Where only one side of the stub connects to the radiating system, as in the J antenna example given here, the radiator length must be exactly right in order to prevent excessive unbalance in the untuned line.

If only one leg of a stub is used to voltage-feed a radiator, it is impossible to secure a perfect balance in the transmission line due to a slight inherent unbalance in the stub itself when one side is left floating. This unbalance, previously discussed under the *Zepp Antenna system,* should not be aggravated by a radiator of improper length.

Current Feed. When a stub is used to current-feed a radiator, the stub should either be left *open* at the bottom end instead of shorted, or else made a *half wave* long. The open stub should be resonated in the same manner as the shorted stub before attaching the transmission line; however, in this case, it is necessary to prune the stub to resonance, as there is no shorting bar.

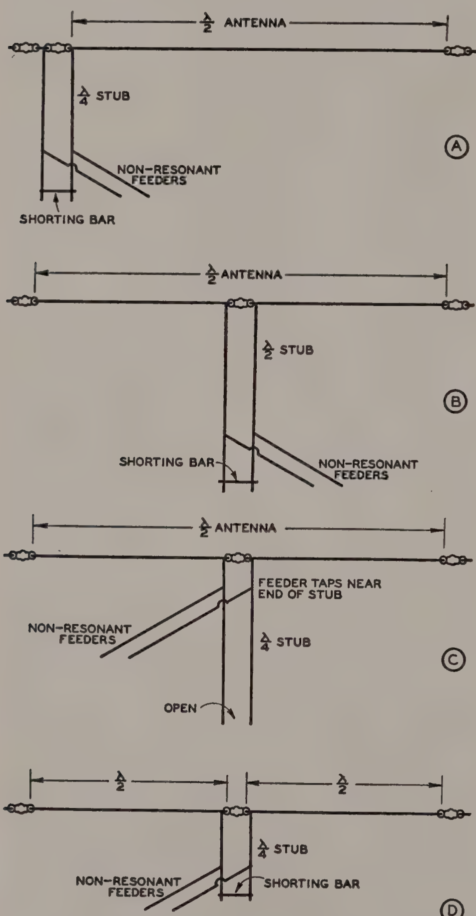


Figure 16.
MATCHING-STUB APPLICATIONS.
(A) Half-wave antenna with quarter-wave matching stub.
(B) Center-fed half-wave antenna with half-wave matching stub.
(C) Center-fed half-wave antenna with stub line cut to exact length without shorting bar.
(D) Two half-wave sections in phase with quarter-wave stub.

Sometimes it is handy to have a stub hang from the radiator to a point that can be reached from the ground, in order to facilitate adjustment of the position of the transmission-line attachment. For this reason, a quarter-wave stub is sometimes made three-quarters wavelength long at the higher frequencies, in order to bring the bottom nearer the ground. Operation with any *odd* number of quarter waves is the same as for the quarter-wave stub.

Any number of *half waves* can be added to either a quarter-wave stub or a half-wave stub without disturbing the operation, though losses will be lowest if the shortest usable stub is employed. This can be fully understood by inspection of the accompanying table.

Stub Length (Electrical)	Current-Fed Radiator	Voltage-Fed Radiator
¼-¾-1¼-etc. wavelengths	Open	Shorted
½-1-1½-2-etc. wave-lengths	Shorted	Open

Shorted-Stub Tuning Procedure. When the antenna requires a shorted stub (odd number of quarter waves if the antenna is voltage-fed, even number of quarter waves if radiator is current-fed), the tuning procedure is as follows:

Shock-excite the radiator (or one of the half-wave sections, if harmonically operated) by means of a makeshift doublet strung directly underneath where possible, and just off the ground a few inches, connected to the transmitter by means of any kind of twisted pair or open line handy.

With the feeders and shorting bar disconnected from the stub, slide along an r.f. milliammeter or low-current dial light at about where you calculate the shorting bar should be, and find the point of maximum current (in other words, use the meter or lamp as a shorting bar).

MAKE SURE IT IS IMPOSSIBLE FOR PLATE VOLTAGE TO BE ON THE FEED LINE BEFORE ATTEMPTING THIS PROCEDURE. *Inductive coupling to the final amplifier by means of a few turns of high tension ignition wire is recommended during any tuning-up process where the operator must come in contact with the antenna or feeders.*

It is best to start with reduced power to the transmitter, until you see how much of an indication you may expect; otherwise, the meter or lamp may be blown on the initial trial. The leads on the lamp or meter should be no longer than necessary to reach across the stub.

After finding the point of maximum current, remove the lamp or meter and connect a piece of wire across the stub at that point.

Starting at a point about a quarter of a quarter wave (8 feet at 40 meters) from the shorting bar, connect the feeders to the stub. Then, move the feeders up and down the stub until the standing waves on the line are at a minimum. The makeshift doublet should, of course, be disconnected, and the regular feeders connected to the transmitter instead during this process. Slight readjustment of the shorting bar will usually result in further improvement.

The standing wave indicator may be either a voltage device, such as a neon bulb, or a current device, such as a r.f. milliammeter connected to a pickup coil. A high degree of accuracy is not required.

The following rule will indicate in which direction the feeders should be moved in an attempt to minimize standing waves: If the current increases on the transmission line as the indicator is moved away from the point of attachment to the stub, the feeders are attached too far from the shorting bar, and must be moved closer to the shorting bar; if the current decreases, the feeders must be attached farther from the shorting bar.

Open-Ended Stub Tuning Procedure. If the antenna requires an open stub (even number of quarter waves if the antenna is voltage-fed, odd number of quarter waves if radiator is current-fed), the tuning procedure is as follows:

Shock-excite the radiator as described for tuning a shorted-stub system, feeders disconnected from the stub, and stub cut slightly longer than the calculated value. Place a field strength meter (the standing wave indicator can be very easily converted into one by addition of a tuned tank) close enough to one end of the radiator to get a reading, and as far as possible from the makeshift exciting antenna. Now, start folding and clipping the stub wires back on themselves a few inches at a time, effectively shortening their length, until you find the peak as registered on the field meter.

Next, attach the feeders to the stub as described for the shorted-stub system, but, for the initial trial connection, the feeders will attach at a distance more nearly three-quarters of a quarter wave from the end of the stub instead of a quarter of a quarter wave, as is the case for a shorted stub. After attaching the feeders, move them along the stub as necessary to minimize standing waves on the line. If sliding the feeders along the stub a few inches makes the standing waves worse, it means the correct connecting point is in the other direction.

Frequency in Kilocycles	Quarter-wave matching section or stub	Half-wave radiator
3500	70′3″	133′7″
3600	68′5″	129′10″
3700	67′6″	126′4″
3800	64′10″	123′
3900	63′1″	119′10″
3950	62′3″	118′4″
4000	61′6″	116′10″
7000	35′1″	66′9″
7150	34′5″	65′4″
7300	33′8″	64′
14,000	17′7″	33′5″
14,200	17′4″	32′11″
14,400	17′1″	32′6″
28,000	8′9″	16′8″
28,500	8′7″	16′5″
29,000	8′6″	16′1″
29,500	8′4″	15′10″
30,000	8′2″	15′7″

DIMENSIONS FOR HALF-WAVE RADIATOR AND QUARTER-WAVE MATCHING STUB OR Q SECTION.

Dimensions for 1750-2000 kc. may be determined from lengths for 3500-4000 kc. A 2000-kc. radiator or matching transformer is just twice as long as one for 4000 kc.

After the optimum point on the stub is found for the feeder attachment, the length of the stub can be "touched up" for a final adjustment to minimize standing waves. This is advisable because the attachment of the feeder often detunes the stub slightly, as will be explained.

Important Note on Stub Adjustment. When a stub is used to match a line to an impedance *of the same order of impedance as that of the surge impedance of the stub and line* (assuming the stub and line use the same wire size and spacing), it will be found that attaching the feeders to the stub introduces a large amount of reactance. The length of the stub then must be altered considerably to restore resonance.

Unfortunately, alteration of the stub length requires that the position of attachment of the feeders be readjusted. Consequently, the

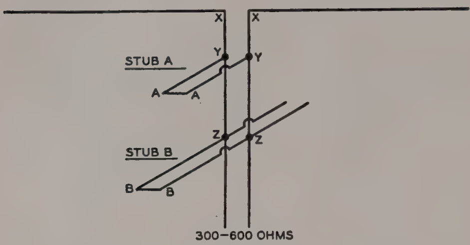


Figure 17.
TWO-FREQUENCY STUB-MATCHED ANTENNA SYSTEM.

Any antenna which has a radiating system capable of efficient operation on two, widely separated frequencies may be matched to an open wire transmission line on both frequencies by use of two "reactance stubs" as shown here. Operation and adjustment are explained in the text.

adjustment entails considerable juggling of both stub length and point of feeder attachment, in order to minimize both reactance and standing waves.

If a *shorted* stub is used to feed an impedance of *more* than 3 times that of the surge impedance of the stub and line, this effect will be negligible, and it is not absolutely necessary that the stub length be readjusted after the feeders are attached. Likewise, the length of an *open* stub need not be altered after attachment of the feeders, if the stub feeds an impedance of *less* than $\frac{1}{3}$ that of the surge impedance of the stub and line.

As a practical example, this means that if a 600-ohm line and shorted stub are used to feed an impedance of more than 1800 ohms, the length of the stub need not be readjusted after the feeders are attached (in order to eliminate objectionable reactance). If the stub feeds an impedance of less than 1800 ohms, attachment of the feeders to the stub will detune the stub appreciably, making readjustment of the stub length absolutely necessary.

When not sure of the exact order of impedance into which the stub works, it is always advisable to try "touching up" the stub length after the feeders are attached.

Two-Frequency Stub Matching. It is practicable to use matching stubs to match an untuned line to an antenna or array on two frequencies. The frequencies need not be harmonically related if the antenna itself is capable of good efficiency on both frequencies. However, the frequencies preferably should be in a ratio not exceeding 4/1 nor less than 1.3/1.

The arrangement is illustrated in Figure 17. The system is tuned up on the lowest frequency for minimum standing waves by means of adjusting the length and point of attachment of stub "A," stub "B" not yet being connected. After the standing waves are reduced to a negligible value, the transmitter is changed to the higher frequency. Stub "B," which is a quarter wave long on the *lower* frequency, then is attached experimentally, and the point of attachment varied until standing waves are at a minimum on the higher frequency. Because stub "B" is exactly a quarter wave long on the lower frequency, its attachment will have virtually no effect upon the operation of the antenna system at the lower frequency.

It should be kept in mind that stub "A" is tuned by varying the distances XY and AY; the stub does not "hang over" as does stub "B." The overall length of stub "B" is not altered; only the distances XZ and BZ are varied when adjusting for minimum standing waves on the higher frequency. It is quite possible that the position of the two stubs will be reversed from that shown in Figure 17. This will depend upon the particular antenna being fed, and the characteristic impedance of the feed line.

Standing Wave Indicators. Many simple devices can be used for detecting the presence and approximate ratio of standing waves on a feed line. A 1-turn pickup loop, about 4 or 5 inches in diameter, may be attached to a current indicator, such as a small Mazda bulb or an r.f. thermogalvanometer, to indicate current excursions along the line. The device should be attached to the end of a wood stick at least a foot long in order to minimize body capacity. The loop is moved along the line in inductive relation to the feed line, care being taken to see that the loop always is in *exactly the same inductive relation* to the line. It should be kept in mind that

CORRECT VALUES OF SURGE IMPEDANCE OF $\lambda/4$ MATCHING SECTIONS FOR DIFFERENT LENGTHS OF ANTENNAS.

Antenna Length in Wavelength	Surge Impedance for Connection Into Two-Wire Open Lines with Impedance of	
	500 Ohms	600 Ohms
½	190	212
1	210	235
2	235	257
4	255	282
8	280	305

Matching section connects into center of a current loop, such as middle of a half-wave section.

this type of indicator is a *current* indicator.

A small neon bulb also may be used to indicate standing waves on a feed line. In this case, the indicator works on *voltage,* and it should be kept in mind that the voltage on the line normally is highest where the current is lowest. This type of indicator is operated by touching various parts of the bulb to *one* feeder wire until an indication of medium brilliancy is obtained. The bulb is then slid along the wire, *in exactly the same position and point of contact with the wire.* If the enamel insulation is not intact on all portions of the wire and the wire is exposed in spots, deceptive "bumps" will be noticed. The wire should be either uniformly insulated or uniformly bare throughout its length; otherwise, it will be necessary to place a thickness of insulating material over the exposed metal parts of the neon bulb, the bulb then working by virtue of capacity to the wire, rather than direct contact.

If it is desired to measure the exact rather than relative standing wave ratio and an r.f. meter is not available, a low range d.c. milliammeter may be used instead, if a suitable rectifier is placed in series with the d.c. meter. A 0–1 ma. d.c. milliammeter in series with a carborundum crystal rectifier is commonly used. As noted before, this type of indicator is a *current* indicator.

Linear R. F. Transformers

Q-Matching Section. A resonant quarter-wave line has the unusual property of acting much as a transformer. Let us take, for example, a section consisting of no. 12 wire spaced 6 inches, which happens to have a surge impedance of 600 ohms. Let the far end be terminated with a pure resistance, and let the near end be fed with radio-frequency energy at the frequency for which the line is a quarter wavelength long. If an impedance measuring set is used to measure the impedance at the near end while the impedance at the far end is varied, an interesting relationship between the 600-ohm characteristic surge impedance of this particular quarter-wave matching line, and the impedance at the two ends will be discovered.

PARALLEL TUBING SURGE IMPEDANCE FOR MATCHING SECTIONS.

Center to Center Spacing in Inches	Impedance in Ohms for ½″ Diameters	Impedance in Ohms for ¼″ Diameters
1	170	250
1.25	188	277
1.5	207	298
1.75	225	318
2	248	335

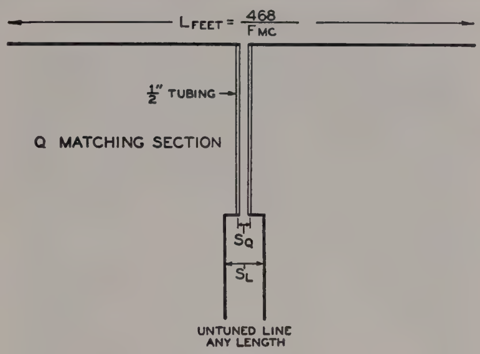


Figure 18.
METHOD OF FEEDING A HALF-WAVE RADIATOR BY MEANS OF Q BARS. REFER TO TABLES FOR DIMENSIONS.

When the impedance at the far end of the line is the same as the characteristic surge impedance of the line itself (600 ohms), the impedance measured at the near end of the quarter-wave line will also be found to be 600 ohms.

Under these conditions, the line would not have any standing waves on it, due to the fact that it is terminated in its characteristic impedance. Now, let the resistance at the far end of the line be doubled, or changed to 1200 ohms. The impedance measured at the near end of the line will be found to have been cut in half, to 300 ohms. If the resistance at the far end is made half the original value of 600 ohms, or 300 ohms, the impedance at the near end doubles the original value of 600 ohms, and becomes 1200 ohms. Therefore, as one resistance goes up, the other goes down proportionately.

It always will be found that the characteristic surge impedance of the quarter-wave matching line is the geometric mean between the impedance at both ends. This relationship is shown by the following formula:

$$Z_{MS} = \sqrt{Z_A Z_L}$$

where
Z_{MS} = Impedance of matching section.
Z_L = Antenna resistance.
Z_A = Line impedance.

Johnson-Q Feed System. The standard form of Johnson-Q feed to a doublet is shown in Figure 18. An impedance match is obtained

	Nº 12 WIRE			Nº 14 WIRE		
	COL. 1	COL. 2	COL. 3	COL. 4	COL. 5	COL. 6
Zo OHMS	SPACING INCHES	SPACING INCHES	CIR. DIA. INCHES	SPACING INCHES	SPACING INCHES	CIR. DIA. INCHES
175	1.415	1 7/16	2.001	1.120	1 1/8	1.585
184	1.495	1 1/2	2.110	1.185	1 3/16	1.675
187	1.535	1 9/16	2.175	1.215	1 1/4	1.720
193	1.630	1 5/8	2.305	1.280	1 5/16	1.820
200	1.720	1 3/4	2.434	1.361	1 3/8	1.935
202 203	1.820	1 13/16	2.560	1.440	1 7/16	2.100
206 207	2.020	2	2.858	1.600	1 5/8	2.261
210 211 212	2.120	2 1/8	3.000	1.630	1 11/16	2.378
216	2.301	2 5/16	3.122	1.825	1 13/16	2.581
219	2.420	2 7/16	3.421	1.920	1 15/16	2.719
223 224 225	2.662	2 11/16	3.700	2.110	1 1/8	2.890
228	2.910	2 15/16	4.110	2.310	2 5/16	3.375
232	3.075	3 1/16	4.350	2.435	2 7/16	3.440
234	3.150	3 1/8	4.450	2.497	2 1/2	3.530
238	3.320	3 5/16	4.690	2.625	2 5/8	3.720
240	3.420	3 7/16	4.835	2.721	2 11/16	3.853
245	3.640	3 5/8	5.150	2.881	2 7/8	4.075
250	4.040	4 1/16	5.710	3.204	3 3/16	4.540
256	4.360	4 3/8	6.160	3.460	3 7/16	4.890
261	4.650	4 5/8	6.580	3.683	3 11/16	5.202

Figure 19.
FOUR-WIRE MATCHING SECTION DESIGN TABLE.

by utilizing a matching section, the surge impedance of which is the geometric mean between the transmission line surge impedance and the radiation resistance of the radiator. A sufficiently good match usually can be obtained by either designing or adjusting the matching section for a dipole to have a surge impedance that is the geometric mean between the line impedance and 72 ohms, the latter being the theoretical radiation resistance of a half-wave doublet either infinitely high or a half wave above a perfect ground.

Though the radiation resistance may depart somewhat from 72 ohms under actual conditions, satisfactory results will be obtained with this assumed value, so long as the dipole radiator is more than a quarter wave above effective earth, and reasonably in the clear. The small degree of standing waves introduced by a slight mismatch will not increase the line losses appreciably, and any *small* amount of reactance present can be tuned out at the transmitter termination with no bad effects. If the reactance is objectionable, it may be minimized by making the untuned line an integral number of quarter waves long.

A Q-matched system can be adjusted precisely, if desired, by constructing a matching section to the calculated dimensions with provision for varying the spacing of the Q section conductors slightly, after the untuned line has been checked for standing waves.

The Q section usually will require about 200 ohms surge impedance when used to match a half-wave doublet, actually varying from about 150 to 250 ohms with different installations. This impedance is difficult to obtain with a two-wire line, as very close spacing would be required. For this reason, either a four-wire line or a line consisting of two half-inch aluminum tubes is ordinarily used. The four-wire section has the advantage of lightness and cheapness, and can be used where the approximate radiation resistance is known with certainty, thus making it possible to design the matching section for a certain value of surge impedance with some assurance that it will turn out to be sufficiently accurate.

The apparent complexity of the Q-matched dipole comes from the large number of antennas and line combinations which the Q section is able to match.

The untuned transmission line between the transmitter and the input, or lower end of the Q section, can be any length (within reason).

Q System with Four-Wire Transformer. The reduction in impedance obtained by the use of 4 conductors instead of 2 makes the four-wire line highly useful for matching transformer applications. For instance, the order of impedance ordinarily required for Q-matching sections is easily obtained by spacing 4 wires around a circular insulating spacer of suitable diameter.

Plastic iced-tea coasters of suitable diameter can be used for spacers. The usual dime store price is 5¢ each. When purchasing the coasters, one should take precaution to get the correct type of material. It seems that some are made from bakelite, while others are made of a plastic that has much better high frequency insulation qualities than bakelite. The plastic ones can easily be identified: they are translucent, while the bakelite ones are not.

The line is flexible, and must be used under slight tension to keep the wires from twisting. Spacers should be placed approximately every 2 feet. The *diagonally opposite* wires should be connected together at each end of the four-wire section.

Exact dimensions for the four-wire type Q section for common surge impedances are given in Figure 19. The length of the section is the same as for the two-conductor type.

OPERATING AN ANTENNA ON ITS HARMONICS

Zepp-fed, single-wire-fed, and direct-fed antennas have always been the most popular antennas for multi-band operation. This is due to the fact that practically all of the antennas that are fed by two-wire nonresonant transmission lines reflect a bad mismatch into the line when operated on 2, 4, or 8 times the fundamental antenna frequency. Thus, the twisted-pair doublet, the Johnson Q, the matched-impedance J or T types, all are unsuitable for even-harmonic operation.

As pointed out earlier in this chapter, the radiating portion of an antenna does not resonate on integral harmonics of its fundamental frequency. Also, if the antenna is several wavelengths long, the point of feed has considerable effect upon the correct length for resonance. The best method of adjusting a harmonically operated antenna is by cut and try. To start with, the antenna can be made an integral number of half wavelengths long, and then pruned as necessary in order to obtain resonance at the desired frequency. A full wave antenna will be found to be about 0.97 wavelength long. The physical and electrical length more nearly correspond as the number of half waves is increased.

When designing an antenna for operation on more than one band, it should be cut for harmonic resonance at its highest operating frequency. If it is to be operated off resonance on some band, it is better to have it off resonance on a low frequency band, because any errors then become a smaller percentage of a half wave.

Dummy Antennas

The law requires the use of some form of dummy antenna when testing a transmitter, in order to minimize unnecessary interference.

The cheapest form of dummy antenna is an electric light globe coupled to the plate tank circuit by means of a 4- to 8-turn pickup coil (or even clipped directly across a few turns of the tank coil). Another good form of dummy antenna that is relatively nonreactive is a bar of carbon tapped across enough of the tank turns to load the amplifier properly. The plaque (noninductive) types of wirewound resistors also are ideal for use as a dummy antenna load.

If a lamp or lamps are chosen of such value that they light up to approximately normal brilliancy at normal transmitter input, the output may be determined with fair accuracy by comparing the brilliancy of the lamps with similar lamps connected to the 110-volt line.

It is difficult to obtain a highly accurate measurement of the output by measuring the r.f. current through the light bulb and applying Ohm's law, because the resistance of the bulb cannot be determined with accuracy. The resistance of a light bulb varies considerably with the amount of current passing through the filament.

For highly accurate measurement of r.f. output, dummy antenna resistors having a resistance that is substantially constant with varying dissipation are offered in 100 watt and 250 watt ratings. These resistors are available in various resistances between 73 and 600 ohms, and can be considered purely resistive at frequencies below 15 Mc. It will be noted that the stock resistance values correspond to the surge impedance of the most common lines. This increases their usefulness.

These resistors are hermetically sealed in glass bulb containers, the latter containing a gas which accelerates the conduction of heat from the resistor element (filament) to the outer surface of the bulb. These resistors glow a dull red at full dissipation rating. Though they somewhat resemble an incandescent lamp physically, they do not produce appreciable light. They may be used in series, parallel, or series parallel to get other resistance values or greater dissipation.

A correction chart is furnished, so that one can correct for the slight non-linearity when a high degree of accuracy is required. With an r.f. ammeter of suitable range in series with the resistor, it is necessary only to note the reading, and refer to the chart to determine the exact amount of power being dissipated in the resistance.

At the higher frequencies (above 30 Mc.), the reactance of the connecting leads, etc. will introduce difficulties, but these may be avoided by resonating the load circuit with a variable series condenser.

COMPACT ANTENNAS

Oftentimes, on the lower frequency bands, it is necessary because of space restrictions, to use a compromise antenna,—an antenna that has been folded or otherwise physically shortened to take up less room than required for a conventional antenna of that frequency. Naturally, a constricted antenna will not have as high efficiency, but, by going about the problem scientifically, it is possible to reduce the size of an antenna considerably with very little sacrifice in efficiency.

Oftentimes, it is possible to get the necessary height for a horizontal antenna on 40, 80, or 160 meters, but not the necessary linear span. As the major portion of the radiation

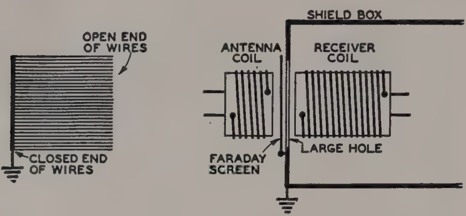


Figure 20
FARADAY ELECTROSTATIC SHIELD.

from a dipole is from the center half of the dipole, the ends may be bent downward with little effect upon the efficiency and radiation pattern. As much as $\frac{1}{8}$ wavelength at each end of a half-wave dipole may be bent or allowed to hang down, if it is necessary in order to get the antenna to fit the span between poles. For the sake of electrical symmetry, it is desirable that the radiator be bent the same amount at each end.

As an example, suppose we should like to string a 130-foot dipole (for 80-meter operation) between two 50-foot poles 90 feet apart. We have 40 feet too much wire; so we shall bend down 20 feet at each end of the dipole. Each bent portion (20 feet) is less than the height of the poles; so there will be no difficulty on that score.

It will be found that when bending a radiator at any point other than a voltage or current loop, the length of the radiator must be increased slightly to restore resonance. Therefore, when bending a half-wave dipole to make it more compact, the length should be increased as necessary to obtain resonance.

Multi-Wire Doublets on Half-Frequency. If we bend down the ends of a half-wave dipole until the bent portion at each end is $\frac{1}{8}$ wavelength long, leaving the flat top $\frac{1}{4}$ wavelength long, we have an antenna of the type just discussed. If we carry the bending process further, and bend the ends not only downwards but back in towards the center, we have something that resembles a multi-wire doublet designed for the next higher frequency band. Thus, we see that a multi-wire half-wave doublet antenna can be used as a *folded* antenna on *half* frequency. The feed line is no longer an untuned feeder, but rather a zepp feed system feeding both ends of the antenna at once. This is possible because the two ends of a dipole are of opposite polarity and phase.

A folded antenna of this type, instead of having very high radiation resistance like a multi-wire doublet system, will have rather low radiation resistance. However, it is still sufficiently high to give good radiation efficiency. The folding of the antenna does not cancel the radiation, because the current is so much greater in the main portion of the antenna than in the ends which are bent in toward the center, and also because the currents in the parallel wires are less than 180 degrees out of phase.

Loading Coils. An old and still popular method of increasing the electrical length of a wire is by means of a *loading coil.* The customary procedure is to place a loading coil at the current loop (center of a dipole or ground lead of a Marconi) and vary the inductance by means of taps until the desired lengthening effect is obtained.

However, the most desirable place for a loading coil is *not* at the *current loop,* but towards the end (voltage loop) of the radiator. If the coil were placed at the extreme end of the antenna, it would have little loading effect, as there would be virtually no current flowing through it. So the coil is placed about $\frac{1}{20}$ wavelength from the end (or ends in the case of a dipole), instead of at the current loop. Thus, we see that while a Marconi will still require only 1 loading coil, a dipole will require 2 for end loading.

As an example of the desirability of end loading, let's look at a vertical Marconi as used in broadcast work. It has been found that a $\frac{1}{8}$-wavelength vertical radiator loaded to an electrical $\frac{1}{4}$ wavelength by means of a loading coil at the bottom or current loop, has a radiation resistance of only 4 or 5 ohms instead of the usual 36 ohms attributed to a $\frac{1}{4}$-wave vertical Marconi.

If we move the loading coil up nearly to the top of the radiator, and add more turns to the coil to compensate for the decreased current flowing through the coil, we find that the antenna now has a radiation resistance of around 20 ohms. Although the height of the radiator is the same, merely by moving the position of the loading coil we have increased the radiation resistance about 5 times.

The exact position of the coil is not critical; approximately $\frac{1}{20}$ wavelength from the far end of a Marconi is a good place for the coil. As previously mentioned, the coil must have considerably more turns to effect resonance than if it were placed at the current loop.

As it is difficult to make adjustments to the coil when it is placed towards the far end of the antenna, the loading coil for an end loaded Marconi is usually wound with somewhat more than the required turns, and resonance found by means of a series condenser in the ground lead. This eliminates the necessity for taking the coil down several times to get precisely the right amount of inductance for resonance at the operating frequency. The series con-

denser also allows one to adjust the antenna for maximum efficiency over the entire band.

The loading coil will be exposed to the weather, and hence this should be taken into consideration. The exact amount of wire required is difficult to calculate, but it will usually be somewhat more than the amount the radiator (including ground lead) lacks of being a quarter wavelength. The coil should be low loss (high Q) for best efficiency. Considerable power will be wasted in a coil having a low Q.

RECEIVING ANTENNAS

A receiving antenna should feed as much signal and as little noise—both man-made and atmospheric—to the receiver as possible. Placing the antenna as high as possible and away from house wiring, etc. will provide *physical* discrimination if a transmission line is used which has no signal pickup. Using a *resonant* antenna will provide *frequency* discrimination, attenuating signals and noise on frequencies removed from the resonant frequency of the antenna. Using a directional antenna will provide *directional* discrimination, attenuating signals and noise reaching the antenna from directions removed from that of the station transmitting the desired signal.

The ideal antenna has these 3 kinds of discrimination: physical, frequency, and directional, which will thus deliver the most signal and the least amount of noise to the input circuit of the receiver. Such an antenna connected to a mediocre receiver will give better results than will the best receiver made, working on a mediocre antenna.

All of the transmitting antennas previously described are suitable for receiving. A good transmitting antenna meets all three of the desirable requirements set forth above. For this reason, an amateur is seldom justified in erecting a separate antenna system for the purpose of receiving. A d.p.d.t. relay designed for r.f. use, working off the send-receive switch or the communications switch on the receiver, can be used to throw whatever transmitting antenna is being used at the time to the receiver input terminals.

Fortunately, the antenna that delivers the best signal into a certain locality will also be best for receiving from that locality, and, conversely, the antenna which provides the best received signal will be best for transmitting to the same locality. In fact, a rotary antenna can be aimed at a station for maximum gain when transmitting by the simple expedient of rotating the array for maximum received signal.

As most man-made noise is essentially vertically polarized, an antenna or array with horizontal polarization will give minimum noise pickup from that source. For this reason, an array with horizontal polarization is advisable when it is to be used not only for transmission but also for reception.

The problem of noise pickup is most important because it is the signal-to-noise ratio that limits the signals capable of being received satisfactorily. No amount of receiver amplification will make a signal readable if the noise reaching the receiver is as loud as the signal. Peak-limiting devices will improve reception when trouble is experienced from *short-pulse* popping noises, such as auto ignition interference. But no electrical device in the receiver is of avail against the steady buzzing, frying noises present in most urban districts.

For the latter type of interference, caused by power leaks, defective neon signs, etc., a recently developed modification of an old principle is oftentimes of considerable help. A noise antenna, a short piece of wire placed so as to pick up as much of the interfering noise and as little of the desired signal as possible, is fed to the input of the receiver *out of phase* with the energy received from the main antenna. By proper adjustment of coupling and experimentation with the length and placement of the noise antenna, it is sometimes possible to eliminate the offending noise completely. The system of noise bucking is described in Chapter 4, *Noise Suppression.*

Stray Pickup. More care has to be taken in coupling a transmission line to a receiver than to a transmitter. The whole antenna system, antenna and transmission line, may tend to act as a Marconi antenna to ground, by virtue of capacity coupling. When transmitting, this effect merely lowers the maximum discrimination of a directive array with but little effect on the power gain; with a nondirectional antenna, nothing will even be noticed when there is a very slight amount of Marconi effect. But if the effect is present when *receiving,* there is little point in using an antenna removed as far as possible from noise sources, because the transmission line itself will pick up the noise.

Faraday Electrostatic Shield. There are two simple ways of avoiding the Marconi effect. The first method calls for a grounded *Faraday screen* between the antenna coil of the receiver and the input grid circuit. This eliminates all capacity coupling. This type of electrostatic screen can be constructed by winding a large number of turns of very small insulated wire on a piece of cardboard which has first been treated with insulating varnish. The wire is wound on, then another coating of varnish is applied.

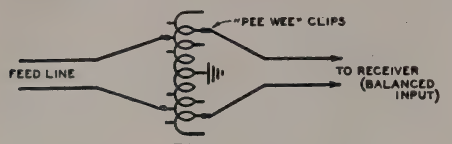


Figure 21.
"AUTOTRANSFORMER" IMPEDANCE MATCHING COIL.
Any two-wire line can be matched to any receiver having balanced input by means of this coupling transformer. The best points at which to tap must be determined by experiment. Both antenna and receiver wires should always be tapped the same number of turns each side of center (ground). A 20- or 25-turn coil wound on a 1-inch form will usually be found optimum. The coil should be tapped every 2 turns, and at the exact center.

After it has dried, *one edge* is trimmed with tin snips or heavy shears, and the wires are soldered together along the opposite edge. The screen is placed between the two coils and grounded. If properly made, it has little effect on the inductive coupling, as there are no closed loops.

Balancing Coils. The second method calls for a center-tapped antenna coil with the center tap grounded. If the coil is not easily accessible, a small center-tapped coil of from 5 to 30 turns is connected across the antenna input to the receiver, and the center tap grounded. While not critical, the best number of turns depends upon the type of transmission line, the frequency, and the turns on the antenna coil in the receiver. For this reason, the correct number of turns can best be determined by experiment.

The center tap must be at the *exact* electrical center of the coil. The coil may be scramble wound, and made self-supporting by means of adhesive tape. It should be borne in mind that a twisted-pair or open two-wire line will work *correctly* only if the receiver has provision for balanced (doublet) input. This is especially true of the latter type of line. If one side of the input or antenna coil is grounded inside the receiver, the ground connection must be broken and moved to the center of the coil, or an external balancing coil used.

Impedance Matching. Another thing to take into consideration is the impedance of the input circuit of the receiver. If the receiver has high impedance input, it will not give maximum performance when a twisted-pair line is used. If it has low impedance input, it will not give maximum performance with an open-wire line. Most receivers are designed with 200- to 300-ohm (medium impedance) input, and will work well with either type line. However, the performance can sometimes be improved by incorporating an impedance matching transformer, even when the receiver has medium impedance (300 ohms) input.

Such a transformer is illustrated in Figure 21. If the line is of lower impedance than the receiver input, the line should be tapped across the fewer number of turns to provide the desired impedance step up. If the line is of higher impedance, the converse applies. Often the coupler will work better if a variable condenser is placed across the entire coil to tune it to resonance.

If the line impedance is lower than that of the receiver, the receiver should be tapped across more turns than the line. If the line impedance is higher than that of the receiver input, the converse applies.

Loop Antennas

As a radiation field contains a magnetic component, it is readily apparent that a coil of wire placed in the proper inductive relation to the magnetic component will serve as an antenna. The efficacy as a pickup antenna is low, as compared to a regular receiving antenna, but because of its compactness and directional characteristics, the loop often is used as a portable antenna, or as a direction indicator.

The loop may be in the form of a circle, square, or rectangle whose length and width are not too widely different. It may be wound in the form of a solenoid, or in the form of a "pancake" helix. For true loop operation, however, the circumference of the loop should not be more than a small fraction of a wavelength.

The loop may be either resonant or nonresonant, though there will be considerable increase in signal pickup when the loop is resonant to the frequency of the signal being received. Also, the directional pattern is different for the two loops, except when both are perfectly balanced to ground, and there is no stray receiver pickup. If there is stray pickup, or the loop is not perfectly balanced, an asymmetrical pattern results *except when the loop is tuned to exact resonance.* With a resonant loop, the only effect of circuit unbalance to ground is to result in the absence of complete nulls; instead, there will be found *minima* as the loop is rotated, the minima being 180 degrees apart, the same as the nulls in a perfectly balanced system.

The result of circuit unbalance to ground, or of stray pickup in the input coupling circuit, permits the whole loop to work against ground as a Marconi antenna. The current thus induced combines with the true loop current. If the loop is resonant, the phasing of

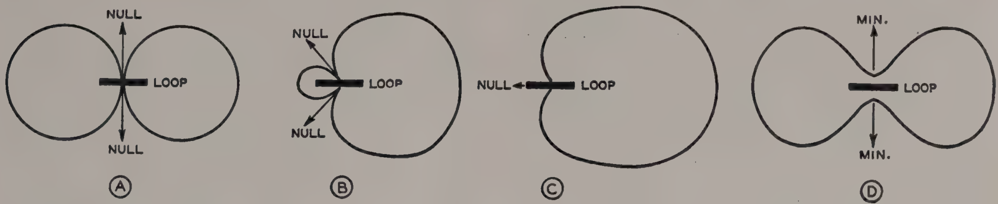


Figure 22.

TYPICAL LOOP ANTENNA PATTERNS.

A: Loop antenna, either resonant or nonresonant, perfectly balanced to ground (no antenna effect).
B: Nonresonant loop antenna, moderate antenna effect.
C: Nonresonant loop antenna, critical amount of antenna effect. Minor lobe completely disappears, leaving only one null.
D: Resonant loop antenna, moderate antenna effect. Nulls are changed to minima, but remain separated exactly 180 degrees.

the two currents is such as to maintain a symmetrical pattern, but there no longer will be complete nulls. If the loop is not resonant, the phasing of the two currents is such as to add in certain directions and cancel completely in others, resulting in an asymmetrical pattern.

Figure 22 shows the patterns obtained under these various conditions. Pattern A is obtained when there is no Marconi effect (also variously known as "antenna effect" or "vertical effect") with either a resonant or nonresonant loop.

With a nonresonant loop, a moderate amount of Marconi effect will produce the pattern shown at B. If the amount of Marconi effect is increased, a point finally will be reached where the small lobe completely disappears, leaving only 1 null. This pattern is shown at C.

A moderate amount of Marconi effect produces the pattern shown at D, when the loop is resonant. When the loop is tuned just slightly off exact resonance, a pattern intermediate between B and D is obtained.

For some applications, the entire loop is enclosed in a static shield. For aircraft work, this shield greatly reduces "rain static." It also virtually eliminates Marconi effect, which is important in the special circuits used in aircraft direction indicators. These instruments give a continuous indication, and have "sense"; that is, they do not have 180 degree ambiguity. However, these instruments are rather complicated, and their theory and operation therefore will not be covered here.

For simple direction finding work, in which two or more bearings are taken, and the station is located by observing the point of intersection on a map, an unshielded resonant loop will be found satisfactory. The only requirement is that the Marconi effect be not too great; otherwise the minima will not be sharply enough defined for accurate bearings.

Loops can be used to take accurate bearings only when the ground wave strongly predominates. When there is appreciable sky wave signal in addition to the ground wave signal, the loop will give inaccurate bearings as a result of downcoming horizontally polarized waves exciting the horizontal portion of the loop when it is adjusted for a null. This is commonly called "night effect," because for certain frequencies it is serious only at night.

While loop antennas can be used for high frequency reception, they are useless as *accurate* direction finders when the signal arrives largely or entirely by sky wave propagation, because sky wave signals do not always follow an exact great circle path.

For microwave and ultra high frequency direction finding, compact beam antennas having a sharp null or maximum are preferable to loop antennas, as loop antennas do not work well at these frequencies. U.h.f. direction finding is discussed in the next chapter.

Practical Direction Finding. In Figure 23 is shown a simple loop and method of connection to the receiver for use on the 160-meter amateur band, or the broadcast band. On these frequencies, bearings accurate to less than 2 degrees can be taken if there is no "night effect," which means 100–200 miles during the day, and 50–75 miles at night. The loop also can be used to provide fair pickup (satisfactory on all except very weak signals) up to about 20 Mc. for determining the *approximate* direction of distant stations, or the exact direction of local stations.

For frequencies below 2000 kc., the loop may be from 1 to 2 feet square, the larger size providing somewhat greater pickup. For frequencies between 2000 and 10,000 kc. it should be about 1 foot square, and above 10,000 kc. about 8 or 10 inches square.

The loop is wound with "bell wire" on a wood frame in the form of a "square solenoid," with an exact even number of turns so that the center tap will come at the bottom of the loop. The tuning condenser C may be an ordinary 350-$\mu\mu$fd. broadcast type, fitted with an insulated shaft extension to minimize body capacity.

A twisted-pair line is used to couple the loop to the receiver, which should have balanced (doublet) input; that is, neither side of the antenna coupling coil should be grounded in the receiver. The twisted line is tapped symmetrically either side of the grounded center tap on the loop, the feed line taps being adjusted together a turn at a time for maximum signal strength.

To take a bearing, simply tune the loop to resonance as indicated by the signal strength meter on the receiver, the loop direction being adjusted roughly for maximum pickup of the signal. Then check to see if the two minima that are observed as the loop is rotated are exactly 180 degrees apart. If not, the loop is not tuned to exact resonance, and the tuning of C should be altered slightly as necessary to cause the two minima or nulls to fall exactly 180 degrees apart. When this is done, either null may be taken as a bearing.

Surrounding metal objects have a tendency to distort the directional pattern of the loop; likewise, large metal objects tend to deflect or reradiate the received signal, resulting in deceptive bearings. To be accurate, loop bearings should be taken with the loop as much in the clear as possible.

Sense Determination. After an accurate bearing is taken with the loop just described, the 180 degree ambiguity can be eliminated as follows.

Tune in a station whose direction is known, and adjust the loop tuning condenser C so that it is considerably on the low capacity side of resonance, as indicated by reduced signal pickup. The pattern then will be similar to B of Figure 22. If the tuning condenser always is tuned to the same side of resonance, and sufficiently off resonance, the small lobe (which is sharper than the large one) will always occur in the direction of the same vertical leg of the loop, which should be given an identifying mark.

Thus, to determine sense, simply detune the condenser C to the low capacity side of resonance, and observe the relative positions of the large and small lobes.

Greatest accuracy will be obtained with a loop located high and in the clear, so that the arriving signal is not disturbed by the presence of surrounding objects, such as steel-frame buildings, telephone lines, etc.

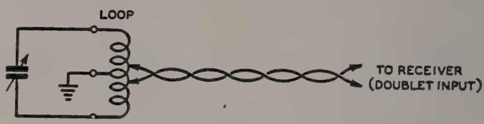


Figure 23.
SIMPLE BUT ACCURATE DIRECTION FINDER.
If the loop is always tuned to resonance, it is not necessary to provide shielding or balancing adjustments in order to obtain accurate readings. For dimensions and data, refer to text.

Supporting the Antenna

The foregoing portion of this chapter has been concerned primarily with the *electrical* characteristics and considerations of antennas. The actual construction of these antennas is just as important. Some of the physical aspects and mechanical problems incident to the actual erection of antennas and arrays will, therefore, be discussed.

Up to 60 feet, there is little point in using mast-type antenna supports unless guy wires must either be eliminated or kept to a minimum. While a little harder to erect, because of their floppy nature, fabricated wood poles of the type to be described will be just as satisfactory as more rigid types, *provided* many guy wires are used.

Rather expensive when purchased through the regular channels, 40- and 50-foot telephone poles *sometimes* can be obtained quite reasonably. In the latter case, they are hard to beat, inasmuch as they require no guying if set in the ground six feet (standard depth), and the resultant pull in any lateral direction is not in excess of a hundred pounds or so.

For heights of 80 to 100 feet, either three-sided or four-sided lattice type masts are most practicable. They *can* be made self-supporting, but a few guys will enable one to use a smaller cross section without danger from high winds. The torque exerted on the base of a high self-supporting mast is terrific during a 40- or 50-mile wind.

Guy Wires. Guy wires should never be pulled taut; a *small* amount of slack is desirable. Galvanized wire, somewhat heavier than seems sufficient for the job, should be used. The heavier wire is a little harder to handle, but costs only a little more and takes longer to rust through. Care should be taken to make sure that no kinks exist when the pole or tower is ready for erection, as the wire will be greatly weakened at such points if a kink is pulled tight, even if it is later straightened.

If "dead men" are used for the guy wire terminations, the wire or rod reaching from

the dead men to the surface should be of non-rusting material, such as brass, or given a heavy coating of asphalt or other protective substance to prevent destructive action by the damp soil. Galvanized iron wire will last only a short time when buried in moist soil.

Only strain-type (compression) insulators should be used for guy wires. Regular ones might be sufficiently strong for the job, but it is not worth taking chances, and egg-type strain insulators are no more expensive.

Only a brass or bronze pulley should be used for the halyard, as a nice high pole with a rusted pulley is truly a sad affair. The bearing of the pulley should be given a few drops of heavy machine oil before the pole or tower is raised. The halyard itself should be of good material, preferably waterproofed. Hemp rope of good quality is better than window sash cord from several standpoints, and is less expensive. Soaking it thoroughly in engine oil of medium viscosity, and then wiping it off with a rag, will not only extend its life but minimize shrinkage in wet weather. Because of the difficulty in replacing a broken halyard (procedure described later), it is a good idea to replace them periodically, without waiting for them to show excessive wear or deterioration.

Screw eyes should not be used in connections where appreciable tension will occur. The bite of the threads is not sufficient to withstand much loading. They should be used only to hold guy wires and such *in position;* the wires should always be wrapped around the mast or pole. Nails will serve just as well, and are cheaper.

Trees as Supports. Often a tall tree can be called upon to support one end of an antenna, but one should not attempt to attach anything to the top, as the swaying of the top of the tree during a heavy wind will complicate matters.

If a tree is utilized for support, provision should be made for keeping the antenna taut without submitting it to the possibility of being severed during a heavy wind. This can be done by the simple expedient of using a pulley and halyard, with weights attached to the lower end of the halyard, to keep the antenna taut. Only sufficient weight to avoid excessive sag in the antenna should be tied to the halyard, as the continual swaying of the tree submits the pulley and halyard to considerable wear.

Galvanized iron pipe, or steel tube conduit, is often used as a vertical radiator, and is quite satisfactory for the purpose. However, when used for supporting antennas, it should be remembered that the grounded supporting poles will distort the field pattern unless spaced some distance from the radiating portion of the antenna.

Painting. The life of a wood mast or pole can be increased several hundred per cent by protecting it from the elements with a coat or two of paint. And, of course, the appearance is greatly enhanced. The wood should first be given a primer coat of flat white outside house paint, which can be thinned down a bit to advantage with second-grade linseed oil. For the second coat, which should not be applied until the first is thoroughly dry, *aluminum paint* is not only the best from a preservative standpoint, but looks very well. This type of paint, when purchased in quantities, is considerably cheaper than might be gathered from the price asked for quarter-pint cans.

Portions of posts or poles below the surface of the soil can be protected from termites and moisture by painting with creosote. While not so strong initially, redwood will deteriorate much more slowly when buried than will the white woods, such as pine.

Antenna Wire. The antenna or array itself presents no especial problem. A few considerations should be borne in mind, however. For instance, soft-drawn copper should not be used, as even a short span will stretch several per cent after whipping around in the wind a few weeks, thus affecting the resonant frequency. Enameled-copper wire, as ordinarily available at radio stores, is usually soft drawn, but by tying one end to some object such as a telephone pole and the other to the frame of an auto, a few husky tugs can be given and the wire, after stretching a bit, is equivalent to hard drawn.

Where a long span of wire is required, or where heavy insulators in the center of the span result in considerable tension, copper-clad steel wire is somewhat better than hard-drawn copper. It is a bit more expensive, though the cost is far from prohibitive. The use of such wire, in conjunction with strain insulators, is advisable, where the antenna would endanger persons or property should it break.

For transmission lines, steel core wire will prove awkward to handle, and hard-drawn copper should, therefore, be used. If the line is long, the strain can be eased by supporting it at several points.

The use of copper tubing for antennas (except at u.h.f.) is not only expensive but unjustifiable. Though it was a fad at one time, there is no excuse for using anything larger than no. 10 copper or copper-clad wire for any power up to 1 kilowatt. In fact, no. 12 will do the trick just as well, and passes the underwriter's rules if copper-clad steel is used. For powers of less than 100 watts, the underwriter's rules permit no. 14 wire of solid copper.

This size is practically as efficient as larger wire, but will not stand the pull that no. 12 or no. 10 will, and the underwriter's rules call for the latter for powers in excess of 100 watts, if solid copper conductor is used.

More important from an electrical standpoint than the actual size of wire used is the soldering of joints, especially at current loops in an antenna of low radiation resistance. In fact, it is good practice to solder *all* joints, thus insuring quiet operation when the antenna is used for receiving.

Insulation. A question that often arises is that of insulation. It depends, of course, upon the r.f. voltage at the point at which the insulator is placed. The r.f. voltage, in turn, depends upon the distance from a current node, and the radiation resistance of the antenna. Radiators having low radiation resistance have very high voltage at the voltage loops; consequently, better than usual insulation is advisable at those points.

Open-wire lines operated as nonresonant lines have little voltage across them; hence, the most inexpensive ceramic types are sufficiently good electrically. With tuned lines, the voltage depends upon the amplitude of the standing waves. If they are very great, the voltage will reach high values at the voltage loops, and the best spacers available are none too good. $\frac{3}{8}$-inch Lucite rod, which can be purchased for 18¢ per foot, permits lightweight spreaders having excellent electrical properties. At the current loops the voltage is quite low, and almost anything will suffice.

When insulators are subject to very high r.f. voltages, they should be cleaned occasionally if in the vicinity of sea water or smoke. Salt scum and soot are not readily dislodged by rain, and when the coating becomes heavy enough, the efficiency of the insulators is greatly impaired.

If a very pretentious installation is to be made, it is wise to check up on both underwriter's rules and local ordinances which might be applicable. If you live anywhere near an airport, and are contemplating a tall pole, it is best to investigate possible regulations and ordinances pertaining to towers in the district, before starting construction.

CHAPTER TWENTY-ONE

Directive Antenna Arrays

No antenna, except a single vertical element, radiates energy equally well in all directions of the compass. All horizontal antennas have directional properties. These usually depend upon the length in wavelengths, the height above ground, and the slope.

The various forms of the half-wave horizontal antenna produce maximum radiation at right angles to the wire, but the directional effect is not great, excepting for very low vertical angles of radiation (such as would be effective on 10 meters). Nearby objects also minimize the directivity of a dipole radiator, so that it hardly seems worth while to go to the trouble to rotate a simple half-wave dipole in an attempt to improve transmission and reception in any direction.

The half-wave doublet, zepp, single-wire-fed, matched impedance, and Johnson Q antenna all have practically the same radiation pattern *when properly built and adjusted.* They all are dipoles, and the feeder system should have no effect on the radiation pattern.

When a multiplicity of radiating dipoles is so located and phased as to reinforce the radiation in certain desired directions and to neutralize radiation in other directions, a directive antenna array is formed.

The function of a directive antenna when used for *transmitting* is to give an increase in signal strength in some direction at the expense of radiation in other directions. For *reception,* one might find useful an antenna giving little or no gain in the direction from which it is desired to receive signals if the antenna is able to *discriminate against interfering* signals and static arriving from other directions. A good directive transmitting antenna, however, generally can also be used to good advantage for reception, as discussed in the previous chapter.

If radiation can be confined to a narrow beam, the signal intensity can be increased a great many times in the desired direction of transmission. This is equivalent to increasing the power output of the transmitter. On the higher frequencies, it is more economical to use a directive antenna than to increase transmitter power, if more than a few watts power is being used.

Directive antennas can be designed to give as high as 23 db gain over that of a single half-wave antenna. However, this high gain (200 times as much power) is confined to such a narrow beam that it can be used only for commercial applications in point-to-point communication.

The increase in radiated power in the desired direction is obtained with a corresponding loss in other directions. Gains of 3 to 10 db seem to be of more practical value for amateur communication, because the angle covered by the beam is wide enough to sweep a fairly large area. Three to 10 db means the equivalent of increasing power from 2 to 10 times.

Horizontal Pattern vs. Vertical Angle. There is a certain optimum vertical angle of radiation for sky wave communication, this angle being dependent upon distance, frequency, time of day, etc. Energy radiated at an angle much lower than this optimum angle is largely lost, while radiation at angles much higher than this optimum angle oftentimes is not nearly so effective.

For this reason, the horizontal directivity pattern as measured on the ground is of no import when dealing with frequencies and distances dependent upon sky wave propagation. It is the horizontal directivity (or gain or discrimination) *measured at the most useful vertical angles of radiation* that is of consequence. The horizontal radiation pattern, as measured on the ground, is considerably different from the pattern obtained at a vertical angle of 15°, and still more different from a pattern obtained at a vertical angle of 30°. In general, a propagation angle of anything less than 30° above the horizon has proved to be effective for 40- and 80-meter operation over long distances. The energy which is radiated at angles higher than approximately 30° above the earth is not very effective at any frequency for extreme dx.

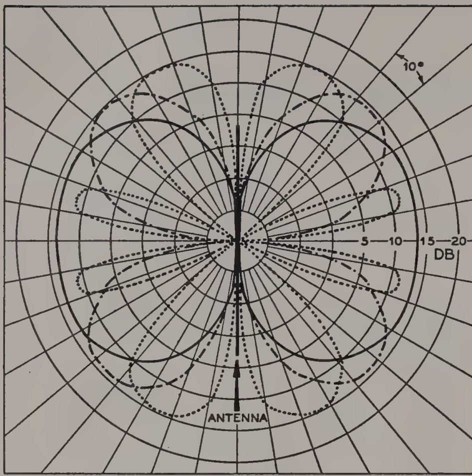


Figure 1.

THEORETICAL FIELD STRENGTH IN DB UNITS FOR THREE TYPES OF ANTENNAS IN FREE SPACE.

To obtain a true picture, one must visualize the radiation lobes in space as encircling the antenna and cutting the page on the dotted lines. The presence of the earth distorts the patterns considerably.

For operation at frequencies in the vicinity of 14 Mc., the most effective angle of radiation is usually about 15° above the horizon, from *any* kind of antenna. The most effective angles for 10-meter operation are those in the vicinity of 10°.

The fact that many simple arrays give considerably more gain at 10 and 20 meters than one would expect from consideration of the horizontal directivity, can be explained by the fact that, besides providing some horizontal directivity, they concentrate the radiation at a lower *vertical* angle. The latter actually may account for the greater portion of the gain obtained by some simple 10-meter arrays. The gain that can be credited to the increased horizontal directivity is never more than 4 or 5 db at most, with the simpler arrays. At 40 and 80 meters, this effect is not so pronounced, most of the gain from an array at these frequencies resulting from the increased horizontal directivity. Thus, a certain type of array may provide 12 to 15 db *effective* gain over a dipole at 10 meters, and only 3 or 4 db gain at 40 meters.

There is an endless variety of directive arrays that give a substantial power gain in the favored direction. However, some are more effective than others taking up the same space; some are easier to feed, and so forth. To include all the various directive antennas that have been developed in the last decade alone would take more space than can be devoted to the subject here.

Long Wire Radiators

Harmonically operated antennas radiate better in certain directions than others, but cannot be considered as having appreciable directivity unless several half wavelengths long. The current in adjoining half-wave elements flows in opposite directions at any instant, and thus, the radiation from the various elements adds in certain directions and neutralizes in others.

A half-wave doublet in free space has a "doughnut" of radiation surrounding it. A full wave has 2; 3 half waves 3; and so on. When the radiator is made more than 4 half wavelengths long, the *end* lobes (cones of radiation) begin to show noticeable power gain over a half-wave doublet, while the broadside lobes get smaller and smaller in amplitude, even though numerous.

The horizontal radiation pattern of such antennas depends upon the vertical angle of radiation being considered. If the wire is more than 4 wavelengths long, the maximum radiation at vertical angles of 15° to 20° (useful for dx) is in line with the wire, being slightly greater a few degrees either side of the wire than directly off the ends. The directivity of the main lobes of radiation is not particularly sharp, and the minor lobes fill in between the main lobes to permit working stations in nearly all directions, though the power radiated broadside to the radiator will not be great if the radiator is more than a few half wavelengths long.

To maintain the out-of-phase condition in adjoining half-wave elements throughout the length of the radiator, it is necessary that a harmonic antenna be fed either at *one end* or at a *current* loop. If fed at a voltage loop, the adjacent sections will be fed *in* phase, and a different radiation pattern will result.

The directivity of a long wire does not increase very much as the length is increased beyond about 15 wavelengths. This is due to the fact that all long-wire antennas are adversely affected by the r.f. resistance of the wire, and because the current amplitude begins to become unequal at different current loops, as a result of attenuation along the wire caused by radiation and losses. As the length is increased, the tuning of the antenna becomes quite broad. In fact, a long wire about 15 waves long is practically aperiodic, and works

LONG-ANTENNA DESIGN CHART
Approximate Length in Feet—End-Fed Antennas

Frequency in Mc.	1λ	1½λ	2λ	2½λ	3λ	3½λ	4λ	4½λ
30	32	48	65	81	97	104	130	146
29	33	50	67	84	101	118	135	152
28	34	52	69	87	104	122	140	157
14.4	66½	100	134	169	203	237	271	305
14.2	67½	102	137	171	206	240	275	310
14.0	68½	103½	139	174	209	244	279	314
7.3	136	206	276	346	416	486	555	625
7.15	136½	207	277	347	417	487	557	627
7.0	137	207½	277½	348	418	488	558	628
4.0	240	362	485	618	730	853	977	1100
3.9	246	372	498	625	750	877	1000	1130
3.8	252	381	511	640	770	900	1030	1160
3.7	259	392	525	658	790	923	1060	1190
3.6	266	403	540	676	812	950	1090	1220
3.5	274	414	555	696	835	977	1120	
2.0	480	725	972	1230	1475			
1.9	504	763	1020	1280				
1.8	532	805	1080					

almost equally well over a wide range of frequencies.

One of the most practical methods of feeding a long-wire antenna is to bring one end of it into the radio room for direct connection to a tuned antenna circuit which is link-coupled to the transmitter. The antenna can be tuned to exact resonance for operation on any harmonic by means of the tuned circuit which is connected to the end of the antenna. This tuned circuit corresponds to an adjustable, non-radiating section of the antenna. A ground is sometimes made to the center of the tuned coil.

If desired, the antenna can be opened and current-fed at a point of maximum *current* by means of a twisted-pair feeder, concentric line, or a Q matching system and open line.

The V Antenna

If two long-wire antennas are built in the form of a V, it is possible to make two of the maximum lobes of one leg shoot in the same direction as two of the maximum lobes of the other leg of the V. The resulting antenna is bidirectional (two opposite directions) for the main lobes of radiation. Each side of the V can be made any odd or even multiple of quarter wavelengths, depending on the method of feeding the apex of the V. The complete system must be a multiple of half waves. If each leg is an even number of quarter waves long, the antenna must be voltage-fed; if an odd number of quarter waves long, current feed must be used.

By choosing the proper angle δ, Figure 2,

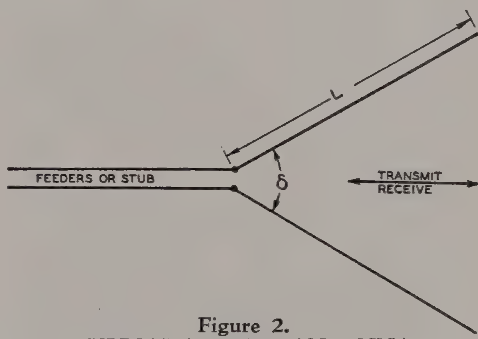


Figure 2.
TYPICAL V-BEAM ANTENNA.

the lobes of radiation from the two long-wire antennas aid each other to form a bidirectional beam. Each wire by itself would have a radiation pattern similar to that for antennas operated on harmonics. The reaction of one upon the other removes two of the four main lobes, and increases the other two in such a way as to form two lobes of still greater magnitude.

The correct wire lengths and the degree of the angle δ are listed in the *V-Antenna Design Table* for various frequencies in the 10-, 20- and 40-meter amateur bands.

The legs of a very long wire V antenna are usually so arranged that the included angle is twice the angle of the major lobe from a single wire if used alone. This arrangement concentrates the radiation of each wire along the bisector of the angle, and permits part of the other lobes to cancel each other.

V-ANTENNA DESIGN TABLE

Frequency in Kilocycles	$L = \lambda$, $\delta = 90°$	$L = 2\lambda$, $\delta = 70°$	$L = 4\lambda$, $\delta = 52°$	$L = 8\lambda$, $\delta = 39°$
28000	34′8″	69′8″	140′	280′
28500	34′1″	68′6″	137′6″	275′
29000	33′6″	67′3″	135′	271′
29500	33′	66′2″	133′	266′
30000	32′5″	65′	131′	262′
14050	69′	139′	279′	558′
14150	68′6″	138′	277′	555′
14250	68′2″	137′	275′	552′
14350	67′7″	136′	273′	548′
7020	138′2″	278′	558′	1120′
7100	136′8″	275′	552′	1106′
7200	134′10″	271′	545′	1090′
7280	133′4″	268′	538′	1078′

With legs shorter than 3 wavelengths, the best directivity and gain are obtained with a somewhat smaller angle than that determined by the lobes. Optimum directivity for a one-wave V is obtained when the angle is 90° rather than 108°, as determined by the ground pattern alone.

If very long wires are used in the V, the angle between the wires is almost unchanged when the length of the wires in wavelengths is altered. However, an error of a few degrees causes a much larger loss in directivity and gain in the case of the longer V than in the shorter one, which is broader.

The V antenna can have each leg either an even or an odd number of quarter waves long. If an even number of quarter waves long, the antenna must be voltage-fed at the apex of the V, while if an odd number of quarter waves long, current feed can be used.

The vertical angle at which the wave is best transmitted or received from a horizontal V antenna depends largely upon the included angle. The sides of the V antenna should be at least a half wavelength above ground; commercial practice dictates a height of approximately a full wavelength above ground.

The Rhombic Antenna

The terminated *rhombic* or *diamond* is probably the most effective directional antenna that is practical for amateur communication. This antenna is non-resonant, with the result that it can be used on three amateur bands, such as 10, 20, and 40 meters. When the antenna is non-resonant, i.e., properly terminated, the system is unidirectional, and the wire dimensions are not critical. The rhombic antenna can be suspended over irregular terrain without greatly affecting its practical operation.

When the free end is terminated with a resistance of a value between 700 and 800 ohms, as shown in Figures 5, 6, and 7, the back-

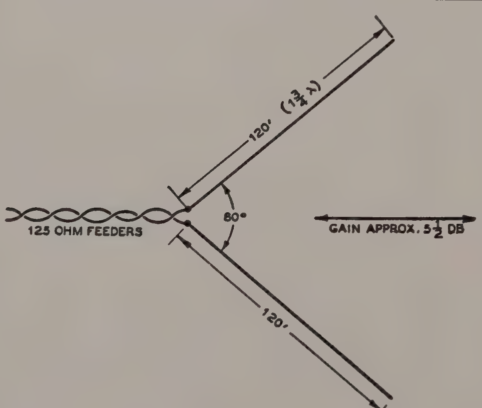


Figure 3.
20-METER V ANTENNA, SMALLEST WORTHWHILE SIZE.

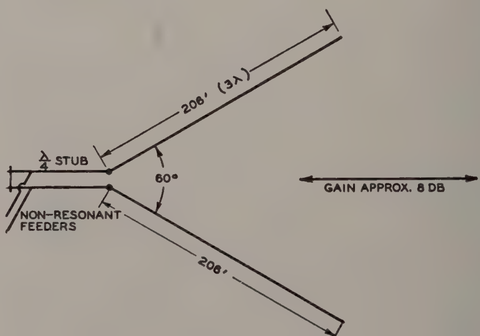
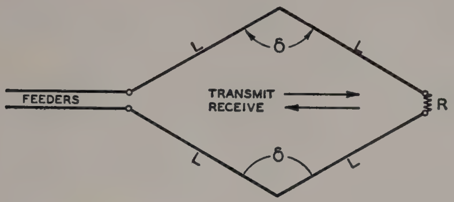
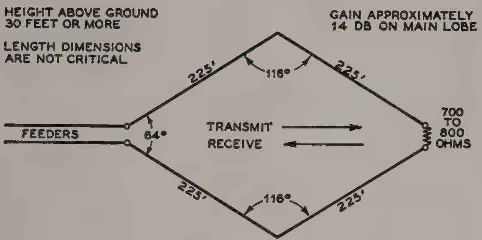


Figure 4.
20-METER V WITH MODERATE GAIN AND DIRECTIVITY.



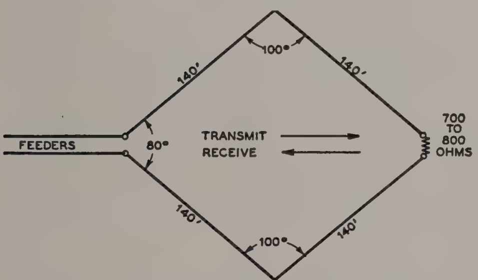
UNTUNED RHOMBIC ANTENNA

Figure 5.



RHOMBIC ANTENNA FOR 7, 14, AND 28 MC. BANDS

Figure 6.



SMALLER RHOMBIC OR DIAMOND ANTENNA
SUITABLE FOR 7, 14, AND 28 MC. BANDS

Figure 7.

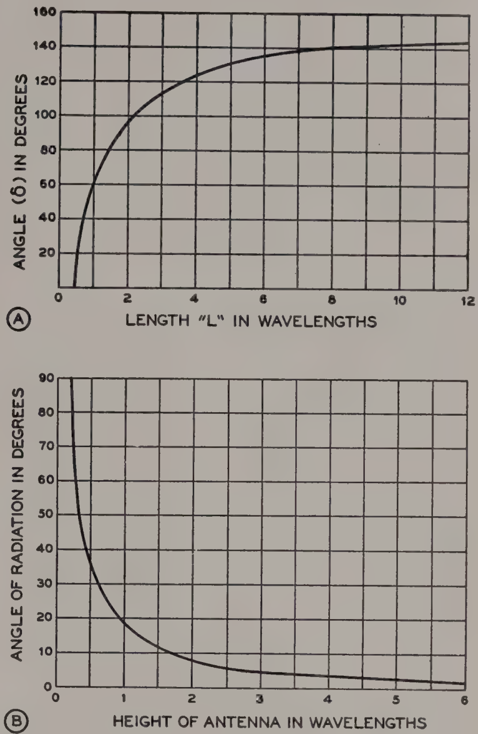


Figure 8.
DIAMOND ANTENNA DESIGN CHARTS.

wave is eliminated, the forward gain is increased, and the antenna can be used on several bands without changes. The terminating resistance should be capable of dissipating one-third the power output of the transmitter, and should have very little reactance. A bank of lamps can be connected in series-parallel for this purpose, or heavy duty carbon rod resistances can be used. For medium or low power transmitters, the non-inductive *plaque* resistors will serve as a satisfactory termination. Several manufacturers offer special resistors suitable for terminating a rhombic antenna.

The terminating device should, for technical reasons, present a small amount of inductive reactance at the point of termination. However, this should not be too great. By using a bank of lamps in series-parallel, this qualification will be met. The total power dissipated by the lamps will be roughly a third of the transmitter output.

Because of the high temperature coefficient of resistance for both carbon and Mazda lamps, neither type is any too satisfactory when used alone, especially in a keyed transmitter. However, by connecting both types in parallel, the resistance can be made fairly constant. This is because the coefficient of one type of lamp is positive, while that of the other is negative. The most constant combination will utilize a 110-volt carbon lamp of 2X watts across each 125-volt Mazda lamp of X watts. Thus, a 60-watt Mazda lamp would have a 120-watt carbon lamp across it. The desired resistance can be obtained by series-connecting or series-paralleling several such units.

A compromise terminating device commonly used consists of a terminated 250-foot or longer length of line, made of resistance wire which *does not have too much resistance per unit length.* If the latter qualification is not

met, the reactance of the line will be excessive. A 250-foot line consisting of no. 25 nichrome wire, spaced 6 inches and terminated with 800 ohms, will serve satisfactorily. Because of the attenuation of the line, the lumped resistance at the end of the line need dissipate but a few watts even when high power is used. A half-dozen 5000-ohm 3-watt carbon resistors in parallel will serve for all except very high power. The attenuating line may be either coiled or folded back on itself to take up less room.

The determination of the best value of terminating resistor must be made while *transmitting,* as the input impedance of the average receiver is considerably lower than 800 ohms. This mismatch will *not* impair the *effectiveness* of the array on *reception,* but as a result, the value of resistor which gives the best directivity on reception will not give the most gain when transmitting. It is preferable to adjust the resistor for maximum gain when transmitting, even though there will be but little difference between the two conditions.

The input resistance of the diamond which is reflected into the transmission line that feeds it is always somewhat less than the terminating resistance, and is around 700 or 750 ohms when the resistor is 800 ohms.

The antenna should be fed with a non-resonant line, preferably with an impedance of approximately 700 ohms. The four corners of the diamond, when possible, should be at least a half wavelength above ground at the lowest frequency of operation. For three-band operation, the proper angle δ for the center band should be observed.

The diamond antenna transmits a horizontally polarized wave at a low angle above the horizon in the case of a large antenna. The angle of radiation above the horizon goes down as the height above ground is increased.

Unless unavoidable, the diamond antenna should not be tilted in any plane. In other words, the poles should be the same height, and the plane of the antenna should be parallel with the ground. Tilting the antenna simply sacrifices about half the directivity, due to the fact that the reflection from the ground does not combine with the incident wave in the desired phase unless the antenna is parallel with the ground.

A good deal of directivity is lost when the terminating resistor is left off and the system is operated as a resonant antenna. If it is desired to reverse the direction of maximum radiation, it is much better practice to run feeders to both ends of the antenna and mount terminating resistors also at both ends. Then, with remote-controlled double-pole double-throw switches located at each end of the antenna, it becomes possible to reverse the array quickly for transmission or reception to or from the opposite direction.

The directive gain of the rhombic antenna is dependent on the height above ground, and the side angle as well as the overall length of each of the 4 radiating wires in the array. Therefore, the gain of a particular array is not easy to calculate.

Stacked Dipole Antennas

The characteristics of a half-wave dipole already have been described. When another dipole is placed in the vicinity and excited either directly or parasitically, the resultant radiation pattern will depend upon the spacing and phase differential, as well as the relative magnitude of the currents. With spacings less than 0.65 wavelength, the radiation is mainly broadside to the 2 wires (bidirectional) when there is no phase difference, and *through* the wires (end fire) when the wires are 180° out of phase. With phase differences between 0° and 180° (45°, 90°, and 135° for instance), the pattern is somewhat unsymmetrical, the radiation being *greater in one direction* than in the opposite direction. In fact, with certain critical spacings, the radiation will be practically unidirectional for phase differences of 45°, 90°, and 135°. However, phase differences of other than 0° and 180° are difficult to obtain except with parasitically excited elements.

With spacings of more than 0.7 wavelength, more than two main lobes appear for all phasing combinations; hence, such spacings are seldom used.

With the dipoles driven so as to be in phase, the most effective spacing is between 0.5 and 0.65 wavelength. The latter provides greater gain, but minor lobes are present which do not appear at 0.5-wavelength spacing. The radiation is broadside to the plane of the wires, and the gain is slightly greater than can be obtained from two dipoles out of phase. The gain falls off rapidly for spacings less than 0.375 wavelength, and there is little point in using spacing of 0.25 wavelength or less with in-phase dipoles, except where it is desirable to increase the radiation resistance. (See *Multi-Wire Doublet.*)

When the dipoles are fed 180° *out of phase,* the directivity is through the plane of the wires, and is greatest with *close spacing,* though there is but little difference in the pattern after the spacing is made less than 0.125 wavelength. The radiation resistance becomes so low for spacings of less than 0.1 wavelength that such spacings are not practicable for antenna arrays except for receiving.

The best *unidirectional* pattern is obtained with 0.1- or 0.125-wavelength spacing, and 135° phase lag. As it is rather difficult to get other than 0° and 180° phasing in driven radiators, parasitic directors and reflectors are usually resorted to for odd values of phasing. These are driven parasitically, rather than directly by feeders, and the phasing can be varied by altering the length of the parasitic elements.

In the three foregoing examples, most of the directivity provided is in a plane at a right angle to the 2 wires, though when out of phase, the directivity is in a line *through* the wires, and when in phase, the directivity is *broadside* to them. Thus, if the wires are oriented vertically, mostly horizontal directivity will be provided. If the wires are oriented horizontally, most of the directivity obtained will be *vertical* directivity.

To increase the sharpness of the directivity in all planes that include one of the wires, additional identical elements are added *in the line of the wires,* and fed so as to be *in phase.* The familiar H array is one array utilizing both types of directivity in the manner prescribed. The 2-section Kraus flat-top beam is another.

These two antennas in their various forms are directional in a horizontal plane, in addition to being low angle radiators, and are perhaps the most practicable of the *bidirectional* stacked-dipole arrays for amateur use. More phased elements can be used to provide greater directivity in planes including one of the radiating elements. The H then becomes a barrage or Sterba array.

For unidirectional work, the most practicable stacked dipole arrays for amateurs are those using close-spaced directors and reflectors (0.1 to 0.125 wavelength spacing).

While there is almost an infinite variety of combinations when it comes to obtaining directivity by means of stacked dipoles, only those systems which are most practical from an amateur standpoint will be discussed at length.

Colinear Antennas

Franklin or *colinear* antennas are widely used by amateurs. The radiation is bidirectional broadside to the antenna. The antenna consists of two or more half-wave radiating sections, with the current *in phase* in each section. This is accomplished by quarter-wave stubs between each radiating section, or by means of a tuned coil and condenser or resonant loading coil between each half-wave radiating section.

Two half waves *in phase* will give a gain of slightly more than 2 db with respect to a single half-wave antenna; three sections will give a gain of approximately 4 db. Additional half-wave sections increase this power gain approximately 1 db per section. The two-section colinear antenna is commonly called a *double zepp.*

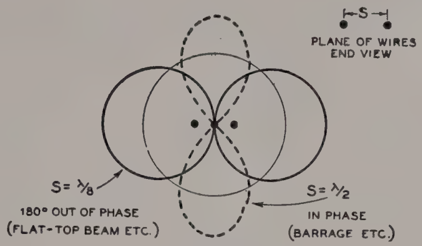


Figure 9.

FIELD STRENGTH PATTERNS OF TWO DIPOLES WHEN IN PHASE AND WHEN OUT OF PHASE.

It can be readily seen that if the dipoles are oriented horizontally most of the directivity will be in the vertical plane; if oriented vertically most of the directivity will be horizontal directivity.

Various feeder systems are shown in the accompanying sketches. A tuned feeder can be used in place of a quarter-wave stub and 600-ohm line. The latter will allow a two-section *colinear* antenna to be operated as a single-section half-wave antenna (current-fed doublet) on half frequency. For example, an antenna of this type would be a half-wave antenna on 40 meters and a 2-section *colinear* antenna on 20 meters. The direction of current at a given instant and the location of the current loops are indicated in the sketches by means of arrowheads and dots, respectively.

Practically all directivity provided by colinear sections is in a horizontal plane. The effect on the vertical directivity is negligible when additional sections are provided. For this reason, the Franklin array is useful particularly on the 40-, 80-, and 160-meter bands, where low angle radiation is not so important. On the higher frequency bands, 20 and 10 meters, an array providing *vertical directivity in addition to horizontal directivity is desirable.* Hence, the Franklin antenna is not as suited for use on the latter two bands as are some of the arrays to be described.

As additional colinear elements are added to a doublet, the radiation resistance goes up much faster than when additional half waves are added out of phase (harmonic operated antenna).

For a colinear array of from 2 to 6 elements, the terminal radiation resistance in ohms at any current loop is approximately 100 times the number of elements.

COLINEAR ANTENNA DESIGN CHART

BAND	FRE-QUENCY IN MC.	L_1	L_2	L_3
10	30	16′	16′5″	8′2″
METERS	29	16′6″	17′	8′6″
	28	17′1″	17′7″	8′8″
20	14.4	33′4″	34′3″	17′1″
METERS	14.2	33′8″	34′7″	17′3″
	14.0	34′1″	35′	17′6″
40	7.3	65′10″	67′6″	33′9″
METERS	7.15	67′	68′8″	34′4″
	7.0	68′5″	70′2″	35′1″
75	4.0	120′	123′	61′6″
METERS	3.9	123′	126′	63′
80	3.6	133′	136′5″	68′2″

It should be borne in mind that the *gain* from a Franklin antenna depends upon the *sharpness* of the horizontal directivity. An array with several colinear elements will give considerable gain, but will cover only a very limited arc.

Double Extended Zepp. The gain of a conventional 2-element Franklin antenna can be increased to a value approaching that obtained from a 3-element Franklin, simply by making the 2 radiating elements 230° long instead of 180° long. The phasing stub is shortened correspondingly to maintain the whole array in resonance. Thus, instead of having 0.5-wavelength elements and a 0.25-wavelength stub, the elements are made 0.64 wavelength long and the stub slightly more than 0.11 wavelength long.

The correct radiator dimensions for a 230° double zepp can be obtained from the *Colinear Antenna Design Chart* simply by multiplying the L_1 values by 1.29. The length for L_3 must be determined experimentally for best results. It will be about $\frac{1}{8}$ wavelength.

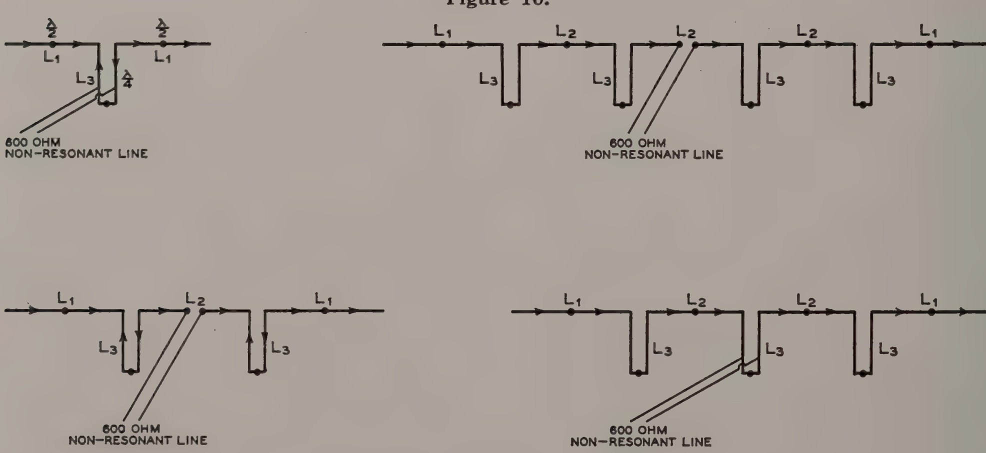
The *vertical* directivity of a colinear antenna having 230° elements is the same as for one having 180° elements. However, parasitic lobes in the horizontal pattern are stronger with the extended version. The radiation resistance of the extended version is slightly lower.

It will be observed that the overall length of the extended zepp, including phasing section, is longer than the 3/2 wavelength wire that makes up a conventional double zepp. The reason for this is that when a wire is bent anywhere except at a voltage or current loop, the wire must be lengthened to restore resonance.

Multiple-Stacked Broadside Arrays

Colinear elements may be stacked above or below another similar string of elements, thus providing vertical directivity. Horizontal colinear elements, stacked two above two and separated by a half wavelength, form the popular "lazy H" array of Figure 11. It is highly recommended for amateur work on 10 and 20 meters when substantial gain without too much directivity is desired. It has high radiation resistance. This results in low voltages and a broad resonance curve, which permits use of inexpensive insulators and enables the array to be used over a fairly wide range

Figure 10.



FRANKLIN OR CO-LINEAR ANTENNAS
HORIZONTAL POLARIZATION BROADSIDE RADIATION

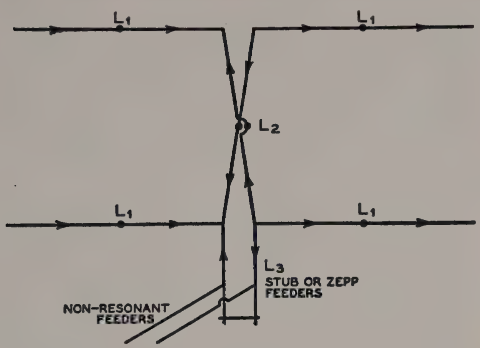
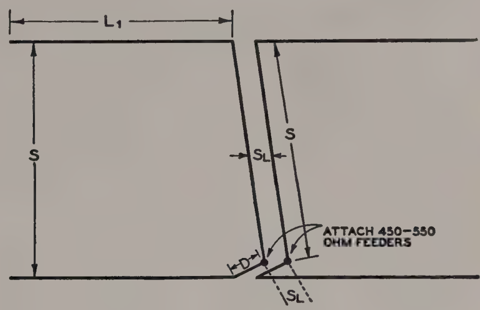


Figure 11.

THE POPULAR "LAZY H" BI-DIRECTIONAL ARRAY.

Stacking the colinear elements results in both vertical and horizontal directivity.



28–29 Mc.	$L_1 = 21\frac{1}{2}'$	$D = 3'$	$S = 24'$
28.5–29.5 Mc.	$L_1 = 21\frac{1}{4}'$	$D = 2\frac{3}{4}'$	$S = 24'$

$S_L =$ $1\frac{1}{2}''$ OR 2" FOR 112 MC.
2" OR 4" FOR 56 MC.
4" FOR 28 MC.
4" OR 6" FOR 14 MC.

THIS ARRAY MAY BE USED ON HALF (NOT TWICE) FREQUENCY WITH GOOD RESULTS. NO CHANGES ARE REQUIRED.

Figure 12.

THE X-H ARRAY, AN EXPANDED VERSION OF THE "LAZY H."

Dimensions for other amateur bands may be determined by multiplying or dividing the specified lengths for the 28-Mc. band by the corresponding figure.

in frequency. For dimensions, see the stacked dipole design table.

The X-H Array. As previously mentioned under the *double extended zepp,* greater horizontal directivity can be obtained from 2 horizontal colinear dipoles by extending each to 230°. It also has been explained previously that cophased dipoles in a curtain arrangement do not show maximum broadside directivity or gain at the common 0.5-wavelength spacing, but at approximately 0.65-wavelength spacing.

Observations of the dimensions of the X-H array in Figure 12 will show that the radiating element lengths have been increased to 230 degrees, and that the spacing has been increased to almost 0.7 wavelength; otherwise, it looks exactly like the familiar lazy H with two exceptions: the phasing section is *not transposed,* and *no matching stub is used.*

Increasing the element lengths and spacing beyond 0.5 wavelength results in stronger parasitic lobes being radiated. However, the magnitude of these lobes is still small, and effects of their presence can be ignored.

The X-H array can be used with good results on *half* (not twice) frequency with no changes whatsoever, thus permitting two-band operation.

The gain at half frequency will not be as great as when the array is used on its regular frequency, but there is still gain over a regular dipole. The general shape of the pattern is the same on both bands, but it will not be so sharp when the array is used on half frequency.

With a line of 450 to 550 ohms, the standing wave ratio (current excursions) will be very low on the higher frequency band, and will be about 2/1 when the array is used at half frequency. If a slight amount of reactance appears at the transmitter end of the line at half frequency, and is found objectionable, the reactance can be eliminated by lengthening or shortening the line as necessary to remove it.

To keep the phasing section from flopping around in the wind, it should be pulled away from the array by means of the feed line. The schematic diagram gives the impression that the phasing section is pulled to one side, because of lack of perspective in such a drawing. In actual practice, the stub should be pulled in such a manner that the phasing section is still at *right angles to the radiating elements.* The feeders need not be pulled tight, but just enough to keep the phasing section from whipping excessively in the wind.

The Sterba "Barrage." Vertical stacking may be applied to strings of colinear elements longer than 2 half waves. In such arrays, the end quarter wave of each string of radiators usually is bent in to meet a similar bent quarter wave from the opposite end radiator. This provides better balance and better coupling between the upper and lower elements when the array is current-fed. Arrays of this type are shown in Figure 13, and are commonly known as Sterba or barrage arrays.

Correct length for the elements and stubs can be determined for any stacked dipole from the *Stacked-Dipole Design Table.*

In these sketches, the arrowheads represent the direction of flow of current at a given instant; the dots represent the points of maxi-

STACKED-DIPOLE DESIGN TABLE

BAND	FRE-QUENCY IN MC.	L_1	L_2	L_3
1.25 METERS	240	24″	24½″	12″
	232	25″	25½″	12½″
	224	26″	26½″	13″
2.5 METERS	120	4′	4′ 1″	24″
	116	4′ 1½″	4′ 3″	25″
	112	4′ 3″	4′ 5″	26″
5 METERS	60	8′	8′ 2″	4′ 1″
	58	8′ 3″	8′ 6″	4′ 3″
	56	8′ 7″	8′ 9″	4′ 5″
10 METERS	30	16′	16′ 5″	8′ 2″
	29	16′ 6″	17′	8′ 6″
	28	17′	17′ 7″	8′ 9″
20 METERS	14.4	33′ 4″	34′ 2″	17′
	14.2	33′ 8″	34′ 7″	17′
	14	34′ 1″	35′	17′ 6″
40 METERS	7.3	65′ 10″	67′ 6″	33′ 9″
	7.0	68′ 2″	70′	35′

mum current and lowest impedance. All arrows should point in the *same direction* in each portion of the radiating sections of an antenna, in order to provide a field *in phase* for broadside radiation. This condition is satisfied by the arrays illustrated in Figure 13.

If 4 or more sections are used in a barrage array, the horizontal directivity will be great enough that the array can be used only over a narrow arc (in 2 opposite directions). For this reason such an array should be oriented with great care.

End-Fire Directivity

By spacing 2 half-wave dipoles, or colinear arrays, at a distance of from 0.1 to 0.25 wavelength and driving the two 180° out of phase, directivity is obtained *through the 2 wires* at right angles to them. Hence, this type of bidirectional array is called *end fire.* A better idea of end-fire directivity can be obtained by referring to Figure 9.

Remember that *end-fire* refers to the radiation with *respect to the 2 wires* in the array, rather than with respect to the array as a whole.

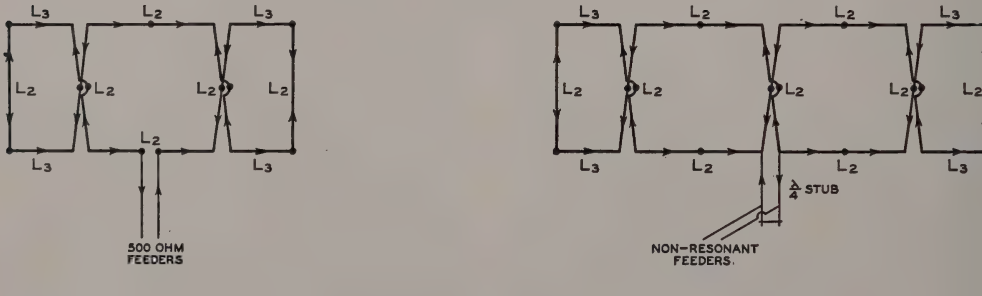
The vertical directivity of an end-fire bidirectional array which is oriented horizontally can be increased by placing a similar end-fire array a half wave below it, and excited in the same phase. Such an array is a combination broadside and end-fire affair. However, most arrays are made either broadside *or* end-fire, rather than a combination of both, though the latter are satisfactory if designed properly.

Kraus Flat-Top Beam. A very effective bidirectional end-fire array is the Kraus *Flat-Top Beam.* Essentially, this antenna consists of 2 close-spaced dipoles or colinear arrays. Because of the close spacing, it is possible to obtain the proper phase relationships in multisection flat tops by crossing the wires at the voltage loops, rather than by resorting to phasing stubs. This greatly simplifies the array. (See Figure 14.) Any number of sections may be used, though the 1- and 2-section arrangements are the most popular. Little extra gain is obtained by using more than 4 sections, and trouble from phase shift may appear.

A center-fed single-section flat-top beam, cut according to the table, can be used quite successfully on its second harmonic, the pattern being similar except that it is a little sharper. The single-section array can also be used on its fourth harmonic with some success, though there then will be four cloverleaf lobes, much the same as with a full-wave antenna.

If a flat-top beam is to be used on more than one band, tuned feeders are necessary.

Figure 13.



HORIZONTALLY POLARIZED BROADSIDE BARRAGE ANTENNAS

The radiation resistance of a flat-top beam is rather low, especially when only one section is used. This means that the voltage will be high at the voltage loops. For this reason, especially good insulators should be used for best results in wet weather.

The exact lengths for the radiating elements are not especially critical, because slight deviations from the correct lengths can be compensated for in the stub or tuned feeders. Proper stub adjustment is covered on page 450. Suitable radiator lengths and approximate stub dimensions are given in the accompanying design table.

Figure 14 shows *top views* of 8 types of flat-top beam antennas. The dimensions for using these antennas on different bands are given in the design table. The 7- and 28-Mc. bands are divided into two parts, but the dimensions for either the low- or high-frequency ends of these bands will be satisfactory for use over the entire band.

In any case, the antennas are tuned to the frequency used, by adjusting the shorting wire on the stub, or tuning the feeders, if no stub is used. The data in the table may be extended to other bands or frequencies by applying the proper factor. Thus, for 56- to 58-Mc. operation, the values for 28 to 29 Mc. are divided by two.

All of the antennas have a bidirectional horizontal pattern on their fundamental frequency. The maximum signal is broadside to the flat top. The single-section type has this pattern on both its fundamental frequency and second harmonic. The other types have 4 main lobes of radiation on the second and higher harmonics. The nominal gains of the different types over a half-wave comparison antenna are as follows: single-section, 4 db; 2-section, 6 db; 3-section, 7 db; 4-section, 8 db.

The current directions on the antennas at any given instant are shown by the arrows on the wires in the figure. The voltage maximum points, where the current reverses phase, are indicated by small X's on the wires.

The maximum spacings given make the beams less critical in their adjustment. Up to one-quarter wave spacing may be used on the fundamental for the 1-section types and also the 2-section center-fed, but it is not desirable to use more than 0.15 wavelength spacing for the other types.

Although the center-fed type of flat top generally is to be preferred because of its symmetry, the end-fed type often is convenient or desirable. For example, when a flat-top beam is used vertically, feeding from the lower end is in most cases more convenient.

If a multisection flat-top array is end-fed instead of center-fed, and tuned feeders are used, stations off the ends of the array can be worked by tying the feeders together and working the whole affair, feeders and all, as a long-wire harmonic antenna. A single-pole double-throw switch can be used for changing the feeders and directivity.

Unidirectional Arrays

If 2 dipoles or colinear arrays are not exactly 0° or 180° out of phase, the pattern becomes unsymmetrical. For certain phasing combinations and spacings, a very good unidirectional pattern is obtained. The required odd values of phasing can be obtained by cutting a parasitically driven element so as to present just the right amount of reactance. Whether the parasitic element acts as a director or reflector depends upon the spacing, and whether the reactance is inductive or capacitive. A parasitic reflector is made just a little longer than an electrical half wavelength, and a director a little shorter than an electrical half wave.

The presence of one or more parasitic elements affects the driven element itself, introducing some reactance, so that slight compensation in the physical length is necessary for resonance. The presence of parasitic elements also reduces the radiation resistance; the more elements, the lower the radiation resistance. Reducing the spacing between the driven dipole and parasitic elements further reduces the radiation resistance. Spacings of 0.1 to 0.125 wavelength are highly satisfactory for either a director or reflector.

The phasing adjustment (length of parasitic elements) is quite critical with respect to frequency, and can best be accomplished by cut and try, and the help of a field strength meter. This is especially true when more than one parasitic element is utilized. It will be found that the adjustment which gives the best forward gain is not the same as that which gives best front to back discrimination, though they are approximately the same.

If only one parasitic element is used, the nose of the directivity pattern will be quite broad, though the front-to-back radiation ratio can be made quite high. The pattern resembles a valentine heart except that the tip is rounded instead of pointed. If the phasing is adjusted for maximum forward gain, rather than maximum discrimination, a small lobe in the backward direction will appear, and the nose of the main lobe will be slightly sharper.

The foregoing applies to the horizontal directivity when the driven and parasitic dipoles are *vertical*. When the dipoles are orientated *horizontally*, the pattern is somewhat different the horizontal directivity *depending upon the*

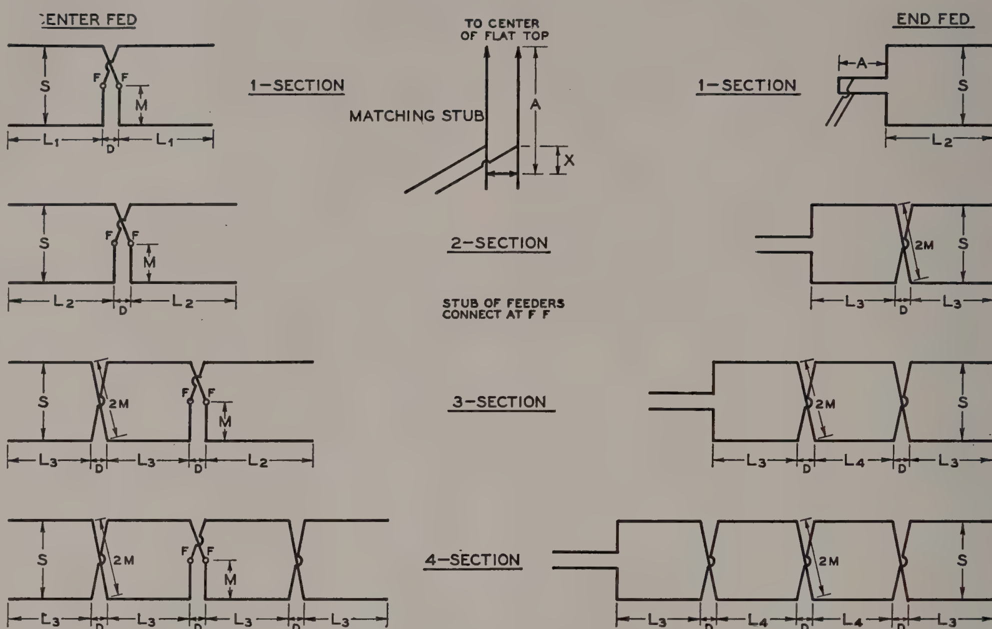


Figure 14.
FLAT-TOP BEAM DESIGN DATA.

FREQUENCY	Spac-ing	S	L_1	L_2	L_3	L_4	M	D	A (1/4) approx.	A (1/2) approx.	A (3/4) approx.	X approx.
7.0-7.2 Mc.	λ/8	17′4″	34′	60′	52′8″	44′	8′10″	4′	26′	60′	96′	4′
7.2-7.3	λ/8	17′0″	33′6″	59′	51′8″	43′1″	8′8″	4′	26′	59′	94′	4′
14.0-14.4	λ/8	8′8″	17′	30′	26′4″	22′	4′5″	2′	13′	30′	48′	2′
14.0-14.4	.15λ	10′5″	17′	30′	25′3″	20′	5′4″	2′	12′	29′	47′	2′
14.0-14.4	.20λ	13′11″	17′	30′	22′10″	.....	7′2″	2′	10′	27′	45′	3′
14.0-14.4	λ/4	17′4″	17′	30′	20′8″	.....	8′10″	2′	8′	25′	43′	4′
28.0-29.0	.15λ	5′2″	8′6″	15′	12′7″	10′	2′8″	1′6″	7′	15′	24′	1′
28.0-29.0	λ/4	8′8″	8′6″	15′	10′4″	.....	4′5″	1′6″	5′	13′	22′	2′
29.0-30.0	.15λ	5′0″	8′3″	14′6″	12′2″	9′8″	2′7″	1′6″	7′	15′	23′	1′
29.0-30.0	λ/4	8′4″	8′3″	14′6″	10′0″	.....	4′4″	1′6″	5′	13′	21′	2′

Dimension chart for flat-top beam antennas. The meanings of the symbols are as follows:

L_1, L_2 L_3 and L_4, the lengths of the sides of the flat-top sections as shown in Figure 37. L_1 is length of the sides of single-section center-fed, L_2 single-section end-fed and 2-section center-fed, L_3 4-section center-fed and end-sections of 4-section end-fed, and L_4 middle sections of 4-section end-fed.

S, the spacing between the flat-top wires.

M, the wire length from the outside to the center of each cross-over.

D, the spacing lengthwise between sections.

A (1/4), the approximate length for a quarter-wave stub.

A (1/2), the approximate length for a half-wave stub.

A (3/4), the approximate length for a three-quarter wave stub.

X, the approximate distance above the shorting wire of the stub for the connection of a 600-ohm line. This distance, as given in the table, is approximately correct only for 2-section flat-tops.

For single-section types it will be smaller and for 3- and 4-section types it will be larger.

The lengths given for a half-wave stub are applicable only to single-section center-fed flat-tops. To be certain of sufficient stub length, it is advisable to make the stub a foot or so longer than shown in the table, especially with the end-fed types. The lengths, A, are measured from the point where the stub connects to the flat-top.

Both the center and end-fed types may be used horizontally. However, where a vertical antenna is desired, the flat-tops can be turned on end. In this case, the end-fed types may be more convenient, feeding from the lower end.

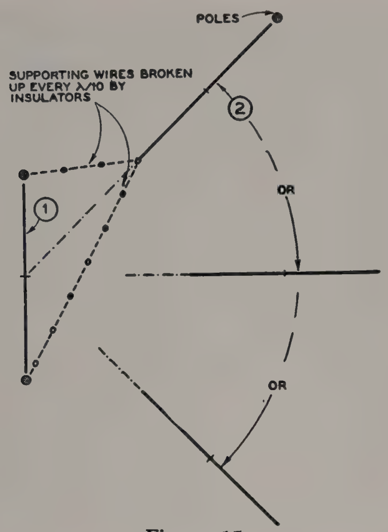


Figure 15.
Illustrating how two dipoles or arrays with horizontal elements can be supported from three poles with a minimum of coupling between the two systems. This is an important consideration if maximum directivity is desired.

vertical angle of radiation. The horizontal directivity is greatest for low vertical angles of radiation when the dipoles are oriented horizontally. For this reason, such an array will exhibit greater discrimination on 10 meters than on 40 meters, for instance.

A close-spaced parasitic director or reflector will lower the radiation resistance of the driven element. If two parasitic elements are used, the radiation resistance will be lowered still more. Consequently, the voltage at the ends of the dipoles of such an array is high, and good insulation is essential, not only because of loss, but because the phasing will be affected by wet weather if poor insulators are used at the high voltage points. Self-supporting quarter-wave rods permit construction of 10- and 20-meter arrays of this type, without the need for insulators at high voltage points.

The low radiation resistance makes the problem of current (center) feed quite difficult. Twisted-pair or concentric line cannot be used without incorporating a matching transformer. A linear transformer of tubing (Q section) cannot be practically designed to have a low enough surge impedance to match a 600-ohm line. A *simple* feed method that is satisfactory is a delta-matched or T-matched open-wire line of from 400 to 600 ohms. The feeder should be attached a short distance each side of the center of the driven dipole. The feeders should be slid back and forth equidistant from the center until standing waves on the line are at a minimum.

A horizontal driven dipole and close-spaced director, or director and reflector, are commonly used as a rotatable array on 10 and 20 meters; such an arrangement is discussed at length later in this chapter.

Orientation of Beam Antennas

Directive antennas, especially those sharp enough to give a large effective power gain, should be so oriented that the line or lines of maximum radiation fall in the desired direction or directions.

To do this, the direction of *true north* must be known with reasonable accuracy. This may vary in the United States by as much as 20 degrees from magnetic north as indicated by a compass.

The magnetic declination (variation of magnetic north in degrees east or west of true north) for any locality in the U. S. A. can be obtained by referring to a map compiled by the U. S. Coast and Geodetic Survey, and available from the Superintendent of Documents, Washington, D. C. The number of the map is 3077, and it is sent only on receipt of 20¢ in *coin.*

A simpler method of determining declination is to inquire of a city engineer or any surveyor or civil engineer in the locality. Any amateur astronomer can help one to determine the direction of true north.

If a beam antenna is to be aimed at a locality more than 2000 miles distant, and the array has a sharp pattern, it will be necessary to use *great circle* directions. A simple method is to stretch a thread from the corresponding two points on a large globe (not a cheap one—they often are inaccurate).

Great circle maps also can be used to determine great circle directions, if such a map is available, centered on a point reasonably close to your locality.

Coupling Between Antennas. If two dipoles or bidirectional arrays are used to cover 4 directions, one will excite the other as a result of electrostatic and electromagnetic coupling, *unless* they are well separated, or care is taken in their orientation. This mutual coupling will result in decreased directivity and a slight loss in gain.

To minimize coupling between two horizontally polarized arrays resonant on the same band, they should be oriented so that a line extended through one of them can be made to intersect the center of the other array. This is illustrated in Figure 15. To eliminate the necessity for 4 poles, antenna no. 2 is sup-

ported at one end by means of a "V" branching out to both of the other 2 poles. These 2 wires should be broken up thoroughly, with insulators every few feet, as they are right in the field of array or antenna no. 1.

ROTATABLE ARRAYS

The radio amateur confined to an apartment house top or a small city lot is at a marked disadvantage when it comes to erecting antennas that will lay down a strong signal at distant points. Even at 10 and 20 meters it is difficult to string up arrays for various points of the compass, without more ground space than is available to the average city amateur. And if the arrays are not placed just right or separated sufficiently, there will be coupling form one array to another, resulting in poor discrimination and directivity. As a result, the city amateur oftentimes turns to a rotatable affair, one which takes up but little ground space and can be aimed in the desired direction.

Unidirectional Rotary Arrays

An effective unidirectional array which is small enough to be rotated without too much difficulty, consists of a horizontal dipole and close-spaced parasitic reflector and director.

The use of 2 parasitic elements instead of 1 adds little to the mechanical difficulties of rotation, and the gain and discrimination (especially the latter) are considerably improved over that obtained with a single director or a single reflector instead of a combination of both. The 3-element array using a close-spaced director, driven element, and close-spaced reflector will exhibit as much as 30 db front-to-back ratio and 20 db front-to-side ratio for *low angle radiation.* The theoretical gain is approximately 8 db over a dipole in free space. In actual practice, the array will usually show 10 db or more gain over a horizontal dipole placed the same height above ground (at 28 and 14 Mc.).

There is little to be gained by using more than 3 elements (one driven and two parasitic). The gain and discrimination are improved very little, and the radiation resistance becomes somewhat low for good efficiency.

There is little to choose as regards the exact spacing of the parasitic elements. Any spacing from 0.1 to 0.15 wavelength may be used for either the director or reflector. However, changes in the spacing will call for slightly different parasitic element lengths.

While the elements may consist of wire supported on a wood framework, self-supporting elements of tubing are to be preferred. The latter type array is easier to construct, looks better, is no more expensive, and avoids the problem of getting sufficiently good insulation at the ends of the elements. The voltages reach such high values towards the ends of the elements that losses will be excessive, unless the insulation is excellent.

Figure 16.
TYPICAL INSTALLATION OF 3-ELEMENT CLOSE-SPACED ARRAY.

The elements may be fabricated of thin-walled steel conduit, hard drawn thin-walled copper tubing, or duralumin tubing. Or, if you prefer, you may purchase tapered copper plated steel tubing elements designed especially for the purpose. Kits are available complete with rotating mechanism and direction indicator, for those who desire to purchase the whole "works" ready to put up.

The radiation resistance of a close-spaced 3-element array is quite low—in the vicinity of 10 ohms. Likewise the Q is high, which means that the array is selective as to frequency. This is perhaps the only important disadvantage of the array; it works much better on the exact frequency for which it was cut, the gain and discrimination falling off considerably either side of resonance.

Because of the high Q and close spacing, it is desirable to use tubing of sufficient diameter that it doesn't whip about appreciably in the wind, as any change in spacing will produce considerable detuning effect.

The self-supporting elements are usually supported on husky standoff insulators, mounted on a wooden cross arm of the type illustrated in Figure 16. The voltage at these points is relatively low, but large insulators are

used for reasons of mechanical strength. The length of each parasitic element is usually made adjustable by means of at least one sliding telescopic joint on either side of center.

The optimum length for the parasitic elements for a given frequency is quite critical, and difficult to predict for a given installation. It will depend upon the type (diameter) and spacing of the elements, primarily, and is best determined by cut and try. For a starter, the reflector may be made exactly $\frac{1}{2}$ wavelength, the driven element 0.96 of a half wavelength, and the director 0.92 of a half wavelength. A half wavelength for a given frequency may be determined by dividing the frequency into 492, the answer being in feet if the frequency is in megacycles.

Set the array temporarily as high above the ground as can be reached conveniently from a ladder or fence. Then, with a local station to give you a check (his receiver must have an "R" meter), adjust the parasitic elements for the best gain. After this point has been found, shorten the director 1 per cent and lengthen the reflector 1 per cent. This improves the discrimination slightly, without reducing the gain appreciably, and makes the array tune more broadly.

Feed Methods. The problem of feeding a 3-element unidirectional array is complicated not only by the problem of rotation, but also by the low radiation resistance. Special low-impedance, flexible coaxial cable with built-in quarter-wave matching section for impedances of this order (10–14 ohms) is available for the purpose, and can be used where the line length is not unduly long. Such cable simply is attached to the center of the driven element, which is split for this type of feed in the same manner as a doublet antenna.

For long line lengths, an open wire line is advisable in the interest of low losses. This type line may be delta matched to the driven element, the same as for a delta matched doublet, except that the points of attachment to the driven element will not be the same as for a simple dipole. The feeder wires are simply fanned out until standing waves are at a minimum. This type of feed does not permit quite as good discrimination, as there is a slight amount of radiation from the fanned out portion of the line, and the director and reflector have little effect on this radiation.

Flexible coaxial line may be allowed to dangle against the supporting tower or guy wires or almost anything without harm, but an open line must be kept from touching anything or twisting on itself and shorting out. This problem often is solved by the incorporation of slip rings and brushes. Not only does this avoid whipping feeders, but permits continuous rotation. Neither voltage nor current is high for a given power at an impedance of 400 to 600 ohms, and there will be little loss in slip rings working at this impedance, if they are carefully constructed.

For 28 Mc. or 56 Mc. operation, the array may be made entirely of pipe, such as thin-walled electrical conduit. With this type of construction, a method of feed is required which does not necessitate "splitting" the driven element in the center. Such an arrangement is illustrated in Figure 17. If the feed line must be longer than about 2 wavelengths

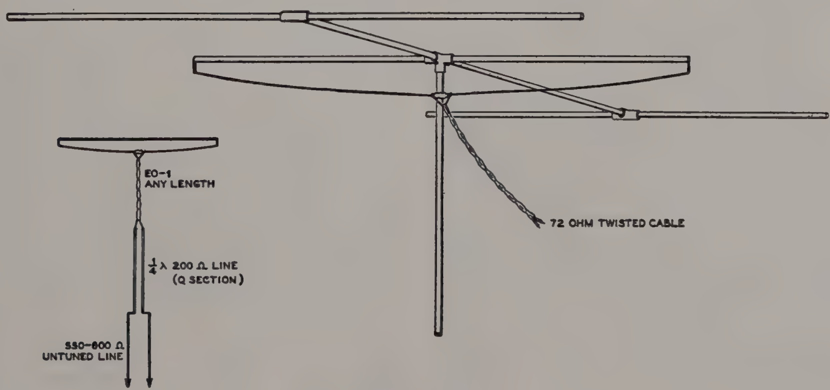


Figure 17.

ALL PIPE VERSION OF 3-ELEMENT CLOSE-SPACED ARRAY.

This type array is widely used on the 28-Mc. amateur band. For long line lengths, an open-wire line is to be preferred in order to minimize losses. The 550-600 ohm line is matched to the 72-ohm cable by means of a quarter-wave "Q" section.

the losses will become appreciable at these frequencies with EO-1 cable, and the arrangement shown in the insert, using a Q section and open wire line, is to be preferred.

The method of feed, shown in the illustration, permits full 360 degree rotation, yet no precautions need be taken with the feed line, as the EO-1 cable can not only touch but even can be wrapped around the supporting pipe without detrimental effects.

With this type feed, the driven element must be slightly longer than for a conventional 3-element array, or about the same length as the reflector. To hold the wire away from the pipe and keep it from slapping the ends of the pipe in a heavy wind, 2 projections are bolted or soldered or brazed to the ends of the pipe to extend downward about 1 or 1½ inches. A piece of no. 12 wire is stretched between these, as shown in the illustration, the wire being split in the center with an insulator.

W8JK Rotary Array

The Kraus "Flat-Top Beam" often is used as a rotary antenna because of its ability to work on two bands when tuned feeders are used. A single-section 14-Mc. flat-top beam can be used as a 2-section 28-Mc. array of the same type. Because the antenna must be fed current on the low-frequency band, and voltage on the high-frequency band, an untuned line is not practicable. Therefore, a tuned line consisting of no. 12 or no. 14 wire spaced 6 inches is advised for two-band operation.

When a tuned line is used, there will be high voltage at the voltage loops, because of the termination mismatch. The line should be designed with this in mind.

The problem of the open line whipping about is simplified by the fact that the array is bidirectional, thus requiring only 180° rotation instead of 360° rotation.

17-foot self-supporting rods are a standard size, and therefore a two-band flat-top beam for 10 and 20 meters is usually made with 34-foot elements, 4 17-foot rods being utilized. The spacing is not critical, 7 or 8 feet being common.

Further details are covered earlier in this chapter under *Flat-Top Beam.* The same considerations apply for a rotatable array as for a stationary one.

Rotating Mechanisms

There are many solutions to the mechanical problem of rotating an array of either of the two types described. The most common system consists of a tower, or else a pole of the "telephone" variety, atop which is an assembly consisting of bearings and driveshaft assembly, the latter supporting a superstructure of wood, which in turn supports the radiating elements.

A simple rotating and drive mechanism for a 10-meter array can be made from a grinding head or saw mandrel, mounted vertically atop the pole or tower. A 20-meter array will require something stronger, an automobile rear axle and housing from a junk dealer serving nicely after being operated on a bit, if the tower or pole will support the weight. The "rear end" of a small car, such as an Austin, is to be preferred to heavier ones.

Another system that has found favor calls for the whole tower's being mounted on a thrust bearing, the entire mast turning inside a large, guyed bearing near the top of the mast.

The cheapest method of rotating the array from the operating position is by means of ropes and pulleys, but motor drive is highly desirable, if one can afford such an installation. Sometimes the motor is placed atop the pole; sometimes it is placed part way down, or at the bottom.

The drive motor must be geared down so that the array turns at a speed of from 1 to 2 r.p.m. The motor and gear reduction assembly from a large oscillating fan can be used to rotate almost any array, as the torque developed by a small motor is quite high, with a gear ratio giving a speed of 1 r.p.m. The oscillating gear shaft on most oscillating fans turns at about 6 r.p.m., and this can be stepped down to 1 or 2 r.p.m. by means of a large and a small pulley, or a bicycle sprocket and chain.

Two Selsyn type motors make the nicest control system, but any small reversible motor is almost as satisfactory. If the feed line to the array is not designed for continuous rotation, an automatic stop should be provided to prevent damage to the feeders.

If pulleys and belt drive are used, the tension of the belt can be adjusted so that it slips on the pulley when the array hits the "stop," thus preventing damage to feeders and motor.

U. H. F. Antennas

THE only difference between the antennas for ultra-high-frequency operation as compared with those for operation in other bands is in their physical size. The fundamental principles are unchanged. For this reason, the reader interested in u.h.f. antennas should first study the discussion of antenna theory in Chapter 20.

Antenna Requirements

Many types of antenna systems can be used for u.h.f. communication. Simple nondirective half-wave vertical antennas are popular for general transmission and reception in all directions. Point-to-point communication is most economically accomplished by means of directional antennas which confine the energy to a narrow beam in the desired direction. If the power is concentrated into a narrow beam, the *apparent* power of the transmitter is increased a great many times.

The useful portion of a signal in the u.h.f. region for short-range communication is that which is radiated in a direction *parallel to the surface of the earth.* A vertical antenna transmits a wave of low angle radiation, and is effective for this reason, not because the radiation is vertically polarized.

Horizontal antennas can be used for receiving, with some reduction in noise. At points close to a transmitter using a vertical antenna, signals will be louder on a vertical receiving antenna. However, at distances far enough from the transmitter that the signal begins to get weak, the transmitted wave has no specific polarization and will appear approximately equal in signal strength on either a vertical or horizontal receiving antenna.

When used for transmitting, horizontal antennas radiate off the ends (in line with the wire) at too high a vertical angle to be effective for quasi-optical u.h.f. work. In fact, even the broadside radiation will be mostly at excessively high angles unless the antenna is far removed from earth (10 or more wavelengths). However, by using several horizontal elements in an array which concentrates the radiation at low angles, results as good or better than with vertically polarized arrays of the same type will be obtained.

The antenna system, for either transmitting or receiving, should be as high above earth as possible, and clear of nearby objects. Transmission lines, consisting of concentric lines or spaced 2-wire lines, can be used to couple the antenna system to the transmitter or receiver. Nonresonant transmission lines are more efficient at these frequencies than those of the resonant type.

Feed Lines. Open lines should preferably be spaced closer than is common for longer wavelengths, as 6 inches is an appreciable fraction of a wavelength at $2\frac{1}{2}$ meters. Radiation from the line will be minimized if $1\frac{1}{2}$-inch spacing is used, rather than the more common 6-inch spacing.

It is possible to construct quite elaborate u.h.f. directive arrays in a small space; even multi-element beams are compact enough to permit rotation. For this reason, it is more common to employ directional arrays to obtain a strong transmitted signal than to resort to high power. Any of the arrays described in Chapter 21 can be used on 5 meters or $2\frac{1}{2}$ meters, though those with sharp, low angle vertical directivity will give the best results. Of the *simpler* types of arrays, those with their dipole elements vertical give the lower angle of radiation, and are to be preferred. When a multi-element stacked dipole curtain is used, little difference is noticed between vertical and horizontal orientation. The angle of radiation is very low in either case.

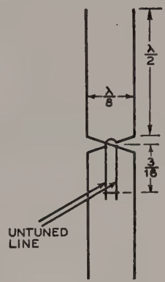
Effect of Feed System on Radiation Angle. A vertical radiator for general coverage u.h.f. use should be made either $\frac{1}{4}$ or $\frac{1}{2}$ wavelength long. Longer antennas do not have their maximum radiation at right angles to the line of the radiator (unless cophased), and, therefore, are not practicable for use where greatest possible radiation parallel to the earth is desired.

Unfortunately, a feed system which is not perfectly balanced and does some radiating,

Figure 1.

U.H.F. W8JK ARRAY ORIENTED FOR VERTICAL POLARIZATION.

For data on this array, refer to preceding chapter. The stub and feed line should be equidistant from the two lower radiating elements.



not only robs the antenna itself of that much power, but *distorts the radiation pattern of the antenna.* As a result, the pattern of a vertical radiator may be so altered that the radiation is bent upwards slightly, and the amount of power leaving the antenna *parallel to the earth* is greatly reduced. A vertical half-wave radiator fed at the bottom by a quarter-wave stub is a good example of this; the slight radiation from the matching section decreases the power radiated parallel to the earth by nearly 10 db.

The only cure is a feed system which does not disturb the radiation pattern of the antenna itself. This means that if a 2-wire line is used, the currents and voltages must be exactly the same (though 180° out of phase) at any point on the feed line. It means that if a concentric feed line is used, there should be no current flowing at all on the outside of the outer conductor.

Means for keeping the feed line out of strong fields where it connects to the radiator are discussed later in the chapter in descriptions of specific antenna systems. The unwanted currents induced in the feed line will be negligible when this precaution is taken.

Radiator Cross Section. In the previous chapter, the statement was made that there is no point in using copper tubing for an antenna (on the medium frequencies). The reason is that considerable tubing would be required, and the cross section still would not be a sufficiently large fraction of a wavelength to improve the antenna characteristics. At ultra high frequencies, however, the radiator length is so short that the expense of large diameter conductors is relatively small, even though copper pipe of 1 inch cross section is used. With such conductors, the antenna will tune much more broadly, and often a broad resonance characteristic is desirable. This is particularly true when an antenna or array is to be used over an entire amateur band.

It should be kept in mind that with such large cross section radiators, the resonant length of the radiator will be somewhat shorter, being only slightly greater than 0.90 of a half wavelength for a dipole when heavy copper pipe is used above 100 Mc.

The matter of using large diameter radiators should not be carried to ridiculous extremes, as detrimental eddy currents will be set up in the conductor. Also, there is little to be gained so far as broadening the resonance characteristic goes after a certain point is reached.

Insulation. The matter of insulation is of prime importance at ultra high frequencies. Many insulators that have very low losses as high as 30 Mc. show up rather poorly at frequencies above 100 Mc. Even the low loss ceramics are none too good where the r.f. voltage is high. One of the best and most practical insulators for use at this frequency is polystyrene (Victron, etc.). It has one disadvantage, however, in that it is subject to cold flow.

It is common practice to so design ultra high frequency antenna systems that the various radiators are supported only at points of relatively low voltage, the best insulation, obviously, being air. The voltages on properly operated *untuned* feed lines are not high, and the question of insulation is not quite so important, though it still should be of good grade.

Polarization. Commercial stations in the U.S.A. favor horizontally polarized antennas for u.h.f. work, both for broadcasting and television. At the present time, however, amateur stations ordinarily use vertically polarized antennas and arrays. As previously mentioned, horizontal polarization results in less noise pickup at the receiver; however, vertical polarization produces greater field strength at relatively short distances.

Horizontally Polarized Arrays

With horizontal antennas and arrays, there is little trouble with undesired currents being induced in the feed line from the field of the radiator. The currents induced in the feed system from the field of one half the antenna or array are cancelled by the currents induced by the field from the other half. The 3-element close-spaced array, the W8JK flat top beam, the lazy-H, and the X-H array will give excellent results at ultra high frequencies when oriented for horizontally polarized radiation, if a 2-wire balanced feeder is used. Dimensions may be determined from the data given in the previous chapter by dividing the specified dimensions by the proper factor for a particular u.h.f. band. The feed line should be spaced somewhat closer than is conventional for lower frequencies. One

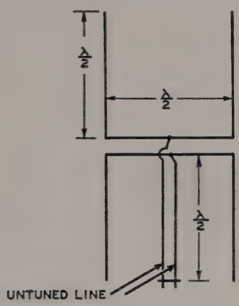


Figure 2.
H TYPE ARRAY ARRANGED FOR VERTICAL POLARIZATION.
The matching stub feeds the center of the phasing section instead of one end as in the case of horizontal orientation. The stub should be equidistant from the two lower radiators.

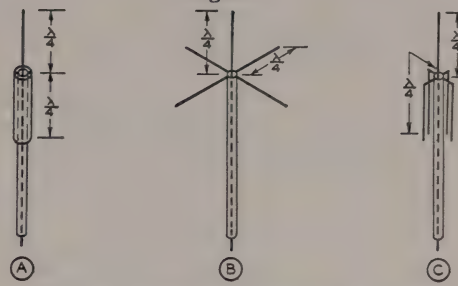


Figure 3.
THREE TYPES OF VERTICAL LOW ANGLE RADIATORS.
At A is shown the "sleeve" type coaxial radiator. The bottom half of the radiator consists of a piece of pipe up through which the coaxial cable runs. At B is illustrated the ground plane vertical, and at C a modification of this antenna.

and one-half inch spacing is recommended for 112 Mc., and either $1\frac{1}{2}$ or 2 inch spacing is satisfactory for 56 Mc.

If large diameter conductors are used as the radiating elements, they must be shortened slightly from the calculated radiator lengths, as the figures given assume ordinary wire radiators.

As 224 Mc. and higher frequencies are not ordinarily used except for short distances, vertical polarization is generally to be preferred above 224 Mc.

Vertically Polarized Antennas and Arrays

Vertical arrays such as the W8JK and the lazy-H (when the latter is fed in the center of the phasing section instead of at one end), will not produce undesired currents in the feed line if a 2-wire feed system is used. Typical examples are shown in Figures 1 and 2. The dimensions refer to electrical length. It is important that the stub and feed line be brought straight down for at least 2 wave lengths; if the stub or line is closer to one radiator than the other, undesired currents will be induced in the feed line.

For general coverage with a single antenna, a single vertical radiator is commonly employed. A 2-wire open transmission line is not suitable for use with this type antenna, and *low loss* concentric feed line is to be recommended. Three practical methods of feeding the radiator with concentric line, with a minimum of current induced in the outside of the line, are shown in Figure 3. Antenna A is known as the "Sleeve" antenna, the lower half of the radiator being a large piece of pipe up through which the concentric feed line is run. At B is shown the Brown ground plane vertical, and at C a modification of this same array. The arrangement shown at B is perhaps the most popular and easiest to construct.

The radiation resistance of the ground plane vertical is approximately 21 ohms, which is not a standard impedance for concentric line. To obtain a good match, the first quarter wavelength of feeder may be of approximately 35 ohms surge impedance, and the remainder of the line of approximately 75 ohms impedance. Thus, the first quarter-wave section of line is used as a matching transformer, and a good match is obtained.

In actual practice the antenna would consist of a quarter-wave rod, mounted by means of insulators atop a pole or pipe mast. Elaborate insulation is not required, as the voltage at the lower end of the quarter-wave radiator is very low. Self-supporting quarter-wave rods would be extended out, as in the illustration, and connected together. As the point of connection is effectively at ground potential, no insulation is required; the horizontal rods may be bolted directly to the supporting pole or mast, even if of metal. The concentric line should be of the low loss type especially designed for u.h.f. use. The outside connects to the junction of the radials, and the inside to the bottom end of the vertical radiator.

Mobile U.H.F. Antennas

For $2\frac{1}{2}$- and 5-meter mobile work, either a quarter wavelength may be used as a Marconi against the car body, or a half wave-

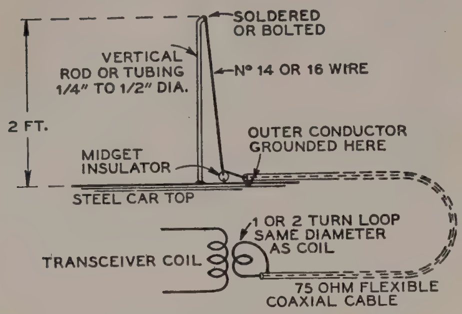


Figure 4.
HIGHLY EFFICIENT 112-MC. MOBILE ANTENNA.
The coaxial cable should preferably be of the type having polystyrene insulation.

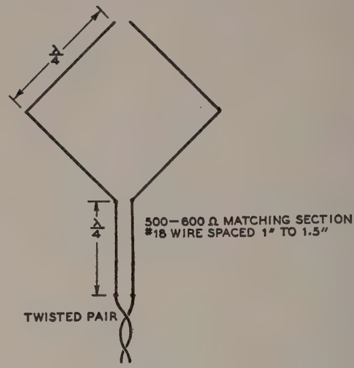


Figure 5.
D.F. ANTENNA SYSTEM FOR U.H.F.
In effect the antenna compares to two vertical close-spaced dipoles 180 degrees out of phase.

length may be used as a vertical dipole. The latter, while delivering a stronger signal, must be very well insulated at the base.

The Marconi type may be fed either with a single wire feeder tapped 28 per cent up from the base, or by means of coaxial line. Coaxial line constructed of copper tubing, with ceramic or polystyrene centering spacers holding the inner conductor, has the lowest loss. If single-wire feed is used, the Marconi antenna need not be insulated at the base. If coaxial line is used, a base insulator is necessary. However, the voltage at the base of a Marconi is quite low, and the insulation need not be especially good.

The coaxial line is connected across the base insulator; no tuning provision need be provided. The radiator length is adjusted for maximum field strength.

Coaxial line may be coupled to the transmitter or receiver by a 1- or 2-turn link.

The losses in *rubber-insulated* coaxial lines are relatively high at u.h.f.; but because only a short length is ordinarily required in a mobile installation, such a line quite often is used when the feeder must be run conveniently and inconspicuously.

An antenna that is highly recommended for 112 Mc. mobile work is illustrated in Figure 4. It consists of a piece of tubing or rod, between ¼ and ½ inch diameter, exactly 2 feet long, mounted vertically just above the center of the windshield atop the car, in about the same position as the auto radio antenna on some of the recent model Ford V-8 cars.

The bottom of the rod or tubing is bolted, welded, or otherwise fastened to the metal portion of the car. The tip of the rod is bent slightly so that when the parallel wire is fastened as shown in the illustration, the wire is held away from the rod sufficiently that it will not whip against the rod as a result of wind or vibration. The wire is anchored by means of a midget insulator, and pulled taut enough that the rod or tubing section bends slightly. Keeping the wire under slight tension will aid in preventing the wire from whipping against the grounded rod or tubing, which would cause the antenna to work erratically.

The outside conductor of the coaxial cable is soldered to the base of the vertical rod, and the inner conductor is soldered to the bottom of the vertical wire where it fastens to the midget insulator.

U.H.F. Direction Finders

For locating a u.h.f. transmitter that is radiating either a horizontally or elliptically polarized wave, a simple horizontal dipole can be used as a direction finder. There will be fairly sharp nulls or minima off the ends of the horizontal radiator. When taking bearings, care must be taken to minimize pickup of reflected waves from surrounding buildings, etc.; otherwise an erroneous bearing will result.

When the polarization of the wave is predominantly or entirely vertical, the antenna illustrated in Figure 5 is recommended. While this array may resemble a loop in mechanical construction, it is a tuned array, and should not be considered as a loop antenna. In effect it is 2 close-spaced dipoles. When the array is turned broadside to the transmitting station, there will be a sharp null. The array is much more sensitive than a conventional loop antenna at u.h.f., and is, therefore, to be preferred. The whole array can be supported from a pole having a single cross arm, which

can be made removable to make the array collapsible.

Microwave D.F. Antennas. Microwave antennas are so small physically that a highly directive array can be contained in a small space. Any highly directive array can be used for direction finding. The arrangement of Figure 5 will be satisfactory up to about 250 Mc. Above this frequency, a dipole equipped with a metal parabolic or angular flat sheet reflector is to be preferred. Such an array is rotated for maximum rather than minimum signal, and has the advantage of providing "sense" even though the accuracy may not be quite as high as that of the various types working on nulls.

The angular flat sheet type reflector (also called "square corner" reflector) is easier to construct, and provides better directivity than a parabolic reflector. One suitable for d.f. work is illustrated in Figure 6. The sheets are the same height as that of the dipole, and about 2 wavelengths on a side. The angle and side length are not critical, but the dipole should bisect the angle accurately. The distance S is approximately 1 wavelength. This array, in addition to being highly directional, provides considerable gain.

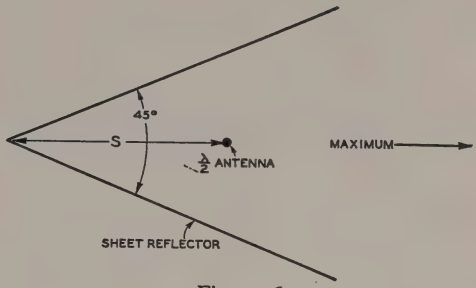


Figure 6.
DIPOLE WITH ANGULAR FLAT SHEET TYPE REFLECTOR.
This array has a very sharp "nose" and may be used for direction finding. It also has considerable gain over a dipole. For dimensions of reflector, refer to text.

The dipole may be fed either with a delta matched open line (with very close spacing), or coaxial cable. The radiation resistance of the dipole will be about 10 ohms.

CHAPTER TWENTY-THREE

Transmitter Adjustment

WHILE there are as many different tuning procedures as there are different types of transmitters, there are certain general rules which should be followed regardless of type. Also, there are certain initial checks that should be made on the transmitter when it is "fired up" for the first time, regardless of type.

Except for very small tubes of the "receiving" classification, it is advisable to permit the filaments to reach operating temperature before applying plate voltage. This takes but a second or two for filament type tubes in the high vacuum class, but about 20 or 30 seconds for heater type tubes and for mercury vapor type rectifiers. It is common practice to allow all filaments to run continuously between consecutive transmissions. Tubes of any type should be allowed to run for at least 15 minutes before applying plate voltage if the tube has not been in service for some time.

In making initial adjustments, it is customary to apply plate voltage to one stage at a time, starting with the oscillator, until the correct tuning adjustments for the whole transmitter have been determined for that frequency or band.

The operation of a crystal oscillator depends to a great extent upon the activity of the crystal, and the activity varies widely with different crystals. The oscillator should be tuned for the greatest output or lowest plate current which will provide strong, stable oscillations. An attempt to adjust the oscillator for every last milliwatt of output will result in the crystal's not starting "cleanly" each time the plate voltage is applied or the key is pressed. A receiver or monitor will be required for this check, during which a check also should be made on the frequency.

The first time the crystal oscillator is operated a check also should be made upon the r.f. crystal current (unless the oscillator is run at very low screen and plate voltages) to make sure that it is not excessive at any setting of the plate tuning condenser.

Tuning of each following stage will depend upon the type of amplifier. However, unless the tube is of the screened grid type or is used only as a doubler, the first thing that should be done is to neutralize the stage correctly. The correct manner in which to neutralize any type of r.f. amplifier is covered thoroughly in Chapter 11, and the reader is referred to that chapter for procedure.

Amplifier stages always should be tuned for maximum output. This does not mean that the coupling must be adjusted until the stage will deliver the maximum power of which it is capable, but that the tank tuning condenser always should be adjusted to the setting which permits maximum output. If the stage is not heavily loaded, this will correspond closely to minimum plate current. However, if the two do not correspond exactly, the stage should be tuned for maximum output rather than minimum plate current. If the difference is appreciable, especially in that amplifier which feeds the antenna, the amplifier should be redesigned to utilize a higher value of tank capacity.

It is natural that the grid current to an amplifier stage should fall off considerably with application of plate voltage, the drop in grid current becoming greater as the loading and plate current on the stage are increased. If the excitation is adjusted for maximum permissible grid current with the tubes loaded, this value will be exceeded when the plate voltage or load is removed, particularly when no grid leak bias is employed. However, under these conditions, the grid impedance drops to such a low value that the high value of grid current represents but little increase in power, and there is little likelihood that the tube will be damaged unless the grid current increases to more than twice its rated maximum.

Screen grid tubes never should be operated with full screen voltage when the plate voltage is removed, as the screen dissipation will become excessive and the tube may be permanently damaged.

When all stages are operating properly, the filament voltage on all tubes should be checked to make sure that it is neither excessive nor deficient, one being about as bad as the other.

Unless the line voltage varies at least several volts throughout the day, filament meters are not required on all stages of a multi-stage transmitter. An initial check when the transmitter is put into operation for the first time is sufficient; after that a single filament meter permanently wired across the filament or filaments of the final amplifier stage will be sufficient. If the filament voltage reads high on that stage, it can be assumed to be high on all stages if the filament voltages were adjusted correctly in the first place. Filament voltage always should be measured *right at the tube socket.*

Parasitic oscillations are capable of causing bad interference to other stations, and a check for them should be made when initial adjustments have been completed. A check for parasitics in a 'phone transmitter should be made with the transmitter being modulated at the full modulation capability of the transmitter, as oftentimes the parasitic will occur only on peaks of the audio cycle. Parasitic oscillations are covered in Chapter 11, and the reader is referred to that chapter. In fact, the whole of Chapter 11 should be read thoroughly before attempting to tune up any transmitter, as an understanding of the considerations involved will make the tuning a relatively simple matter. In the case of a 'phone transmitter, the reader is referred also to Chapter 8 for amplitude modulation and 9 for frequency modulation adjustments.

Antenna Coupling

When coupling either an antenna or antenna feed system to a transmitter, the important considerations are as follows: (1) means should be provided for varying the load on the amplifier, (2) the two tubes in a push-pull amplifier should be equally loaded, (3) the load presented to the final amplifier should be nonreactive; in other words, it should be a purely resistive load.

The first item is often referred to as "matching the feeder impedance to the transmitter" or "matching the impedance." It is really a matter of *loading*. The coupling is increased until the final amplifier draws the correct plate current. Actually, all the matching and mismatching we worry about pertains to the junction of the feeders and *antenna.*

The matter of equal load on push-pull tubes can be taken care of by simply making sure that the coupling system is symmetrical, both physically and electrically. For instance, it is not the best practice to connect a single-wire feeder directly to the tank coil of a push-pull amplifier.

The third consideration, that of obtaining a nonreactive load, is important from the standpoint of efficiency, radiated harmonics, and voice quality in the case of a 'phone transmitter. If the feeders are clipped directly on the amplifier plate tank coil, either the surge impedance of the feeders must match the antenna impedance perfectly (thus avoiding standing waves) or else the feeders must be cut to exact resonance.

If an inductively-coupled auxiliary tank is used as an antenna tuner for the purpose of adjusting load and tuning out any reactance, one need not worry about feeder length or complete absence of standing waves. For this reason, it is always the safest procedure to use such an antenna coupler rather than connect directly to the plate tank coil.

Function of an Antenna Coupler. The function of an output coupler is to transform the impedance of the feed line, or the antenna, into that value of plate load impedance which will allow the final amplifier to operate most effectively. The antenna coupler is, therefore, primarily an impedance transformer. It may serve a secondary purpose in filtering out harmonics of the carrier frequency. It may also tune the antenna system.

Practically every known antenna coupler can be made to give good results when properly adjusted. Certain types are more convenient to use than others, and the only general rule to follow in the choice of an antenna coupler is to use the simplest one that will serve your particular problem.

There is practically nothing that an operator can do at the station end of a transmission line that will either increase or decrease the standing waves on the line, as that is entirely a matter of the coupling between the line and the antenna itself. However, the coupling at the station end of the transmission line has a very marked effect on the efficiency and the power output of the final amplifier in the transmitter. Whenever we adjust antenna coupling and thus vary the d.c. plate current on the final amplifier, all we do is vary the ratio of impedance transformation between the feed line and the tube plate (or plates).

Coupling Methods. Figure 1 shows several of the most common methods of coupling between final amplifier and feed line.

The fixed condenser C_B is a large capacity mica condenser in every case. It has no effect upon tuning or operation; it is merely a blocking condenser keeping high voltage d.c. off the transmission line.

Capacitive Coupling. Figure 1A shows a simple method of coupling a single-wire nonresonant feeder to an unsplit plate tank. The coupling is increased by moving the tap away from the voltage node and toward the plate end of the plate tank coil. Either the center

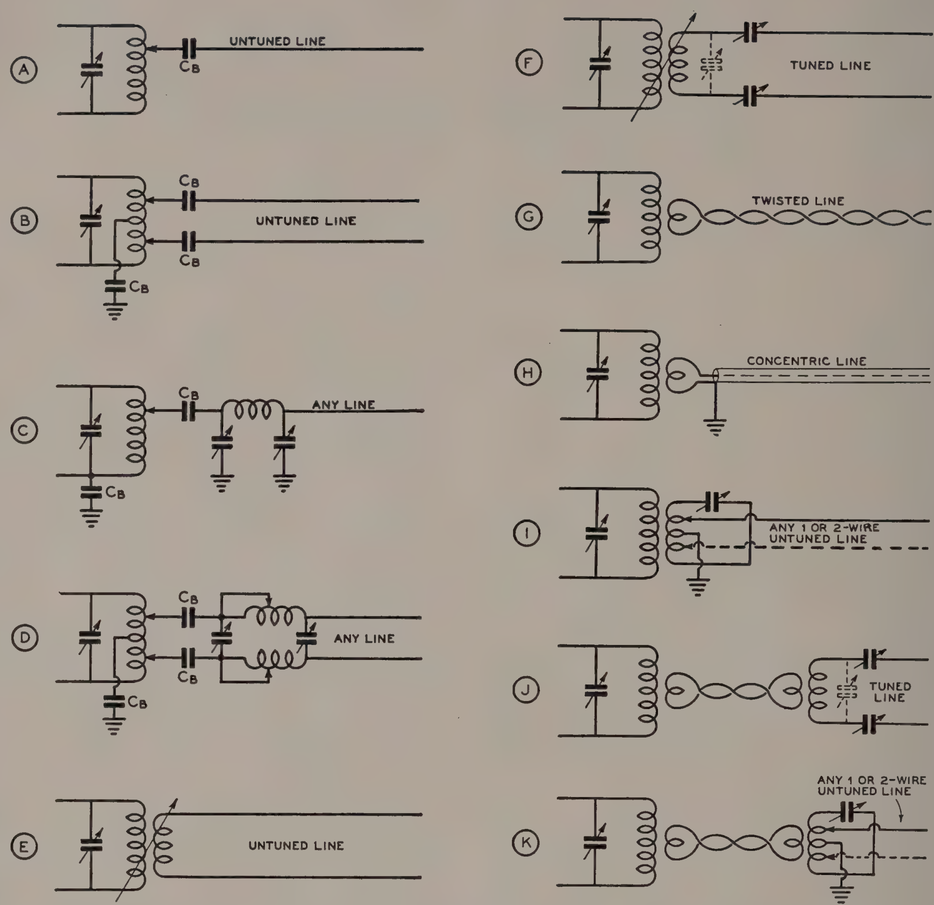


Figure 1.

COMMON METHODS OF COUPLING TRANSMISSION LINES TO THE OUTPUT TANK OF THE TRANSMITTER.

Balanced 2-wire lines are assumed, whether of the resonant or "flat" (untuned) type. Coupling turns should always be placed around the "cold" portion of the coil; whether this is the center or end will be determined by whether the coil has one end grounded or is balanced to ground (center at ground r.f. potential). Tank tuning condensers can be split stator where balanced tanks are shown (center at low r.f. potential) without affecting operation of coupling circuit. C_b indicates mica blocking condenser to keep d.c. plate voltage off the feeder; these condensers should have a working voltage in excess of peak plate voltage and be at least .001 μfd.

or the bottom end of the coil may be by-passed to ground.

The system shown in Figure 1B shows a means of coupling an untuned 2-wire line to a split plate tank. If it is desired to couple a 2-wire untuned line to an unsplit plate tank, it will be necessary to use some form of inductive coupling. See Figure 1E.

The circuit of Figure 1C shows a π-section filter coupling an unsplit tank to any end-fed antenna or single-wire line. Figure 1D shows the 2-wire version of the π-section coupler, sometimes called the *Collins* coupler.

Inductive Coupling. Inductive coupling methods may be classified in two types: direct inductive coupling and link coupling. Direct inductive coupling has been very popular for years, but link coupling between the plate tank and the antenna coupler proper is usually more desirable. Figure 1E shows inductive coupling to an untuned 2-wire line. This same arrangement can be used to couple from a split plate

tank to a single-wire untuned feeder by grounding one side of the antenna coil.

The circuit shown in Figure 1F is the conventional method of coupling a zepp or tuned feed line to a plate tank circuit, but the arrangement shown in Figure 1J is easier to adjust. Circuit shown in Figure 1I is for coupling either a single or 2-wire untuned feeder to either a split or unsplit plate tank circuit. The arrangement shown in Figure 1K is easier to adjust. All coupling links anywhere in a transmitter should be coupled at a point of low r.f. potential to avoid undesired capacitive coupling.

Untuned low impedance lines of the twisted pair and coaxial types can best be coupled inductively by means of a 1- or 2-turn coupling link around the plate tank coil at the voltage node.

Tuning Pi-Section Filter. To get good results from the π-section antenna coupler, certain precautions must be followed. The ratio of impedance transformation in π networks depends on the ratio in capacity of the two condensers C_1 and C_2 (Figure 1C and D).

The first step in tuning is to disconnect the π-section coupler from the plate tank entirely. Then, apply low plate voltage and tune the plate tank condenser to resonance. Remove the plate voltage and tap the π-section connection or connections approximately half-way between the cold point on the coil and the plate or plates. Adjust C_2 to approximately half maximum capacity and apply plate voltage. Quickly adjust C_1 to the point where the d.c. plate current dips, indicating resonance.

At the minimum point in this plate current dip, the plate current will either be higher or lower than normal for the final amplifier. If it is lower, it indicates that the coupling is too loose; in other words, there is too high a ratio of impedance transformation. The plate current can be increased by *reducing* the capacity of C_2 and then restoring resonance with condenser C_1. At no time after the π-section coupler is attached to the plate tank should the plate tuning condenser be touched. If the d.c. plate current with C_1 tuned to resonance is too high, it may be reduced by *increasing* the capacity of C_2 in small steps, each time restoring resonance with condenser C_1.

Should the plate current persist in being too high even with C_2 at maximum capacity, it indicates either that C_2 has too low maximum capacity, or that the π-section filter input is tapped too close to the plate of the final amplifier. If the plate current *cannot* be made to go high enough even with condenser C_2 at minimum capacity, it indicates that the input of the π-section is not tapped close enough to the plate end of the plate tank coil.

Mechanical Considerations. If inductive coupling to the final amplifier is contemplated, attention must be given to the mechanical or physical considerations. Variable coupling is a desirable feature which facilitates correct loading of the amplifier. It is more easily incorporated if but a few turns are involved. This explains the popularity of link coupling methods (such as Figure 1K) over directly coupled systems of the type illustrated in Figure 1I. Untuned lines of 600 ohms or less, when operating correctly, seldom require more than a half dozen turns in the coupling link to provide sufficient coupling, especially on the higher frequency bands. Twisted-pair lines or coaxial cable may require only 1 or 2 turns. Marconi antennas (no feed line) may require anywhere from 1 to 10 turns, depending upon the frequency and radiation resistance.

Because sometimes the next integral turn provides too much coupling while without it there is insufficient coupling, it is necessary to provide means for obtaining coupling intermediate between that provided by integral turns. This can be done by adding the next integral turn and then either pulling the coupling coil away from the tank coil a little, or enlarging the turns so that the coupling coil does not fit snugly over the tank coil.

One very satisfactory method of providing continuously variable coupling calls for a set of split tank coils, with 1- or 1½-inch spacing between the two halves of the coils (depending upon diameter of the coils). A swinging coupling link, with sufficient tension or friction on the hinge to maintain the link in position after it has been adjusted, can be inserted between the two halves of the tank coil to give any degree of coupling desired. Manufactured coils can be obtained with this system of adjustable coupling. Another type manufactured coil is wound on a ceramic coil form with individual link turning inside the form on a shaft supported on bearings inserted in the form. The latter type requires two extra contacts on the coil jack bar.

If one uses the simpler method of pushing coupling turns down between the turns of the tank coil until sufficient coupling is obtained, high tension ignition cable is recommended if the plate voltage of more than 500 volts appears on the plate tank coil. Hookup wire or house wire is satisfactory for lower voltages.

The coupling link should never be placed at a point of high voltage on the tank coil. This means that the coupling link should be placed around the *center* of a *split* plate tank or near the "cold" *end* of an *unsplit* tank coil.

For a given number of turns in the coupling link, greatest coupling will occur when the link is placed around the center of the coil, regard-

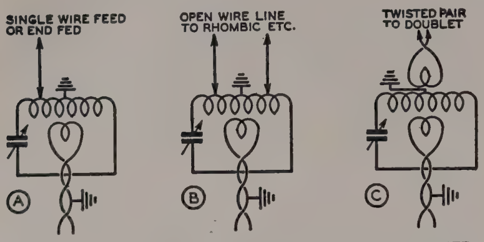


Figure 2.
SIMPLE METHODS OF HARMONIC SUPPRESSION WITH AN AUXILIARY TANK CIRCUIT.

less of the location of the node on the coil. For this reason, it is sometimes difficult to get sufficient coupling with an unsplit tank, as the link must be placed at the cold end of the coil in such a system to prevent detuning of the tank circuit, possible arcing between tank coil and link, and capacity coupling of harmonics.

On the higher frequencies, it is important that superfluous reactance is not coupled into the line by a pick-up link having an excessive number of turns. This means that instead of using a 10 turn link on 28 Mc. to couple to a 72-ohm line and backing off on the coupling coil until the desired coupling is obtained, the number of turns should be reduced and the pick-up coil coupled tighter to the tank coil. For this reason, it is difficult to construct a swinging-link assembly having a single multi-turn coupling coil for coupling on all bands. With this type coupling, it usually will be found that if the pick-up coil has sufficient turns to permit optimum coupling on 160 meters, the coil will be so large that it will couple in an objectionable amount of reactance at 28 Mc. This assumes that the transmitter works into a line of the same surge impedance on all bands.

Suppressing Harmonic Radiation. Harmonics are present in the output of nearly all transmitters, though some transmitters are worse offenders in this respect than others. Those that are strong enough to be bothersome are usually the second and third harmonics.

Current-fed antennas, such as the twisted-pair-fed doublet and the Johnson Q-fed doublet, discriminate against radiation of the even harmonics. This is what keeps these antennas from being used effectively as all-band antennas. However, they *are* responsive to the *odd* harmonics, working about as well on the third harmonic as on the fundamental. For this reason, any third harmonic energy present in the output of the transmitter will be radiated unless a harmonic trap is used or other means taken to prevent it.

Most all-band antennas are responsive to both odd and even harmonics, and therefore are still worse as regards the possibility of harmonic radiation.

The delta-matched antenna, and radiators fed by means of a shorted stub and untuned line, provide about the best discrimination against harmonics, but even these will radiate some third and other odd harmonic energy.

Best practice indicates the reduction of the amount of harmonic component in the transmitter output to as low a value as possible, then further attenuation between the transmitter and antenna *regardless of what antenna and feed system is used.*

Three definite conditions must exist in the transmitter before harmonic radiation can take place. First, the final amplifier must either be generating or amplifying the undesired harmonics; second, the coupling system between the amplifier and the feeders or antenna system must be capable of either radiating them or transmitting them to the antenna, and third, the antenna system (or its feeders) must be capable of radiating this harmonic energy.

One effective method of reducing capacity coupling is through the use of a Faraday shield. The Faraday shield, however, offers no attenuation to anything but *capacity coupling* of the undesired energy. Since a great deal of the harmonic energy (the third and other odd harmonics) is *inductively* coupled to the antenna system, an arrangement which will attenuate both capacitively and inductively coupled harmonics (both odd and even) would be desirable. A Faraday shield is not a cure-all. However, its performance is effective enough to warrant inclusion as standard equipment.

A simple and very effective method of harmonic suppression is shown in Figure 2. The link from the final tank to the antenna tank should consist of either a length of low impedance cable (EO1 or similar) or a *closely spaced* ($\frac{1}{2}$ inch) line of no. 12 or larger wire. This link should be loosely coupled by means of a single turn on 10, 20, and 40 meters (2 turns on 80 and 160 meters) at either end to both tank circuits. One side of the link should be effectively grounded near the final tank.

The antenna tank itself should be of medium C (a Q of about 10 or 12) at the operating frequency. At Figure 2C the two links, the one to the final and the one to the antenna, should be spaced about 2 inches or so apart and at equal distances from the grounded center of the antenna coil. The balance of the diagram should be self-explanatory.

This coupling system operates by virtue of the fact that capacity coupling between the

final tank and the antenna is eliminated by the grounded link and the grounded center tap of the antenna tank; also, due to the selectivity of the antenna tank against the harmonic frequencies, inductive coupling of them into the antenna system will also be attenuated.

In closing, a few general "don'ts" might be in order:

Don't use two tubes in parallel. Put them in push-pull if possible.

Don't use a doubler to feed an antenna unless it is of the push-push type. In a single-ended doubler, there is a high percentage of ½ and 2X output frequency present in the output tank.

Don't use more bias and excitation than necessary for reasonable efficiency or (in a 'phone transmitter) good linearity.

Don't use a 75-meter zepp on 160 meters, a 40-meter zepp on 80 meters, etc. Although it is usually the odd harmonics that are inductively coupled, in this case the second harmonic will be inductively coupled and elimination of capacity coupling will not remove the second harmonic.

Don't use an all-band antenna unless you do not have room for separate antennas. If you must use such an antenna, use a harmonic-attenuating tank as shown in the accompanying diagrams.

Run a test with some local station close enough to give you an accurate check, and see if your harmonics are objectionable.

A Simple Universal Coupler. A split-stator condenser of 200 $\mu\mu$fd. or more per section can be mounted on a small board along with a large and a small multitapped coil to make a very useful and versatile antenna coupler and harmonic suppressor. With this unit it is possible to resonate and load almost any conceivable form of radiator and tuned feed system, and to adjust the loading and provide harmonic suppression with almost any untuned transmission line. The circuit is shown in Figure 3.

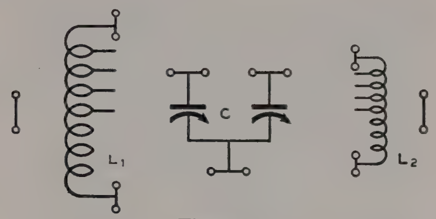
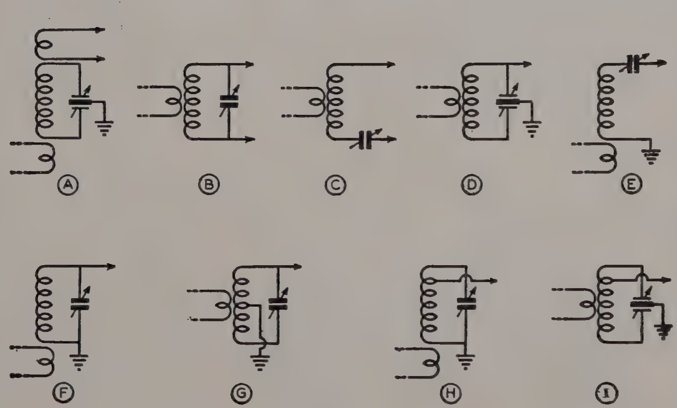


Figure 3.
CIRCUIT DIAGRAM OF THE UNIVERSAL COUPLER.
The dots indicate heavy Fahnestock clips. For coil and condenser constants, see text.

To facilitate connecting the coil and condenser combination in the many different ways possible, 12 large-size dual Fahnestock connectors are mounted on the coils and condenser terminals and generously scattered around. Two are mounted on standoff insulators to act as terminals for ground, antenna, or other wires. A dozen lengths of heavy flexible wire of random lengths between 6 and 18 inches enable one to connect up the components in an almost infinite variety of combinations. Low-voltage auto ignition cable or heavy flexible hookup wire will do nicely.

Because under certain conditions and in certain uses both rotor and stator will be hot with r.f. voltage, an insulated extension is provided for the condenser shaft in order to remove the dial from the condenser by a few inches. This effectively reduces body capacity. It also precludes the possibility of being burned by the dial set-screw.

Figure 4.
APPLICATIONS OF THE UNIVERSAL ANTENNA COUPLER.



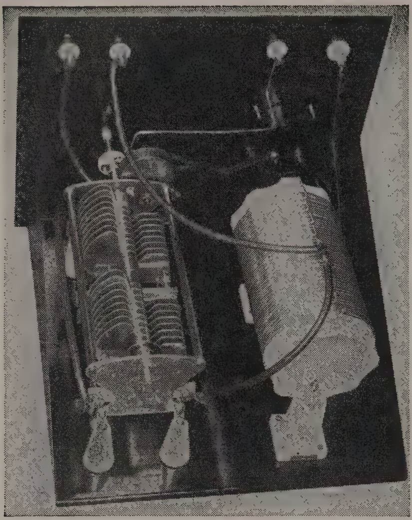


Figure 5.
UNIVERSAL ANTENNA COUPLER.
The unit shown above was designed for use only on the two lower frequency bands, hence the smaller coil (L_2 of Figure 3) was omitted. An r.f. ammeter has been added for convenience in tuning.

The large coil consists of 30 turns of no. 12 wire, 4 inches in diameter and spaced to occupy $5\frac{3}{4}$ inches of winding space. The small coil consists of 14 turns, 2 inches in diameter, spaced to occupy $3\frac{1}{4}$ inches of winding space. Heavy duty 80- and 20-meter coils of commercial manufacture will serve nicely.

Both coils have taps brought out every other turn from one end to the center to facilitate clipping to the coils. A copper or brass clip is preferable to a steel clip for shorting out turns as the circulating current may be quite high.

Now to cover some of the things one can do with this simple contraption:

At A in Figure 4 the unit is used as a harmonic suppression tank as advocated earlier in this chapter.

Combination B may be used for either an end-fed or center-fed zepp that requires parallel tuning. It may also be used to feed an untuned open line, providing harmonic suppression. It may be used to tune an antenna counterpoise system that has a higher natural frequency than that upon which it is desired to operate. It may be used with any system utilizing tuned feeders where the system cannot be resonated with series tuning (see C).

Combination C may be used for either an end-fed or center-fed zepp that requires series tuning. It may be used to feed an antenna counterpoise system that is too long electrically to resonate at the operating frequency at its natural period. It may be used with a multiband antenna where the feeders are too long. It may be used for almost any system utilizing tuned feeders where the system cannot be resonated using B.

Arrangement D may be used for feeding an end-fed antenna (even number of quarter waves long). It is usually preferable to F which is sometimes used for the same purpose.

System F also is used to tune a Marconi that is slightly shorter than an odd number of quarter waves long.

System E is the common method for tuning a Marconi where the antenna is slightly longer than an odd number of quarter waves.

G is commonly used to end-feed an antenna an even number of quarter waves long. It is a variation of D.

H and I are used for feeding either a single-wire-fed antenna or for end-feeding a very long-wire antenna (6 or more wavelengths long). For the latter purpose these are preferable to D, F, and G.

In each case the link is coupled around the coil being used, and one side of the twisted pair feeding the link is grounded. Coaxial cable may be substituted for the twisted pair if the builder so desires.

Test and Measuring Equipment

EVERY radio station and laboratory should possess certain pieces of test equipment, in order to insure proper operation of radio receivers, transmitters, amplifiers and antenna systems, and to diagnose trouble when it occurs. Other pieces of test equipment, while very convenient and undoubtedly desirable, are not absolutely necessary and may be considered somewhat of a luxury.

A simple volt-ohmmeter, an absorption wavemeter, and monitor may be considered essential equipment. The last can be designed and calibrated to act as a frequency meter. How much additional test equipment an amateur is justified in acquiring depends upon the condition of his pocketbook, the amount of money otherwise invested in his station, and his ingenuity and resourcefulness.

Some amateurs can diagnose trouble and determine whether their equipment is operating properly by means of a single meter, a few resistors and various parts from the junk box, though the job would undoubtedly be facilitated by a more extensive array of test equipment. Other amateurs, particularly those less technically inclined, will find more need for various special-purpose test instruments when trouble hunting or tuning up a transmitter; in fact, they usually will be helpless without such instruments.

Virtually everything in the way of test equipment of use to the amateur is described in this chapter. The units have been designed with simplicity and economy in mind, but not at the expense of reasonable accuracy or versatility.

Volt-Ohmmeters

Perhaps the most useful single piece of test equipment is the *volt-ohmmeter*. The circuit of a typical volt-ohmmeter of the multiple range type is shown in Figure 1. Tap 1 is used to permit use of the instrument as a 0-1 ma. milliammeter. Tap 2 permits accurate reading of resistors up to 100,000 ohms; taps 3, 4, 5, and 6 are for making voltage measurements, the full scale voltages being 10, 50, 250, and 500 volts respectively.

The 1000-ohm potentiometer is used to bring the needle to zero ohms when the terminals are shorted; this adjustment should always be made before a resistance measurement is taken.

Higher voltages than 500 can be read if a higher value of multiplier resistor is added to an additional tap on the switch. The proper value for a given full scale reading can be determined from Ohm's law.

Resistances higher than 100,000 ohms cannot be measured accurately with the circuit constants shown; however, by increasing the ohmmeter battery to 45 volts and multiplying the 4000-ohm resistor and 1000-ohm potentiometer by 10, the ohms scale also will be multiplied by 10. This would permit accurate measurements up to 1 megohm.

0-1 ma. meters are available with special volt-ohmmeter scales which make individual calibration unnecessary. Or, special scales can be purchased separately and substituted for the original scale on the milliammeter.

Obviously the accuracy of the instrument either as a voltmeter or as an ammeter can be no better than the accuracy of the milliammeter and the resistors.

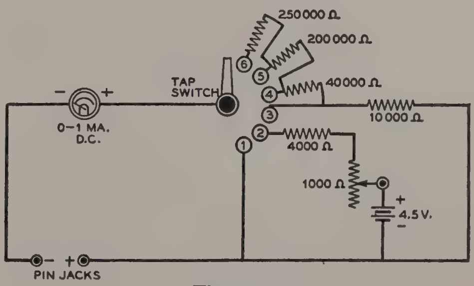


Figure 1.

VOLT-OHMMETER CIRCUIT.

Position 1 of Switch.....0-1 ma.
Position 2 of Switch.....0-100,000 ohms
Position 3 of Switch.....0-10 volts
Position 4 of Switch.....0-50 volts
Position 5 of Switch.....0-250 volts
Position 6 of Switch.....0-500 volts

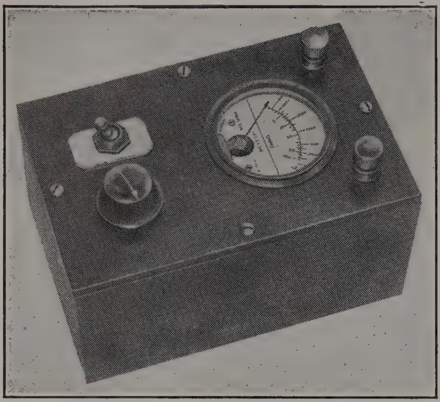


Figure 2.
LOW RANGE OHMMETER.
This ohmmeter is particularly useful for measuring resistances too low to be read accurately on an ohmmeter of the type illustrated in Figure 1.

Because volt-ohmmeters are so widely used and because the circuit is standardized to a considerable extent, it is possible to purchase a factory-built volt-ohmmeter for no more than the component parts would cost if purchased individually. For this reason no construction details are given. However, anyone already possessing a suitable milliammeter and desirous of incorporating it in a simple volt-ohmmeter should be able to build one from the schematic diagram and design data given here. Special, precision (accurately calibrated) multiplier resistors are available if extreme accuracy is desired.

Medium- and Low-Range Ohmmeter

Most ohmmeters, including the one just described, are not adapted for accurate measurement of low resistances—in the neighborhood of 100 ohms, for instance.

The ohmmeter illustrated in Figure 2 was especially designed for the reasonably accurate reading of resistances all the way down to 1 ohm. Two scales are provided, one going in one direction and the other scale going in the other direction because of the different manner in which the milliammeter is used in each case. The low scale covers from 1 to 100 ohms and the high scale from 100 ohms to 10,000 ohms. The high scale is in reality a medium-range scale. For accurate reading of resistances over 10,000 ohms, an ohmmeter of the type previously described should be used.

The 1-100 ohm scale is useful for checking transformers, chokes, r.f. coils, etc., which often have a resistance of only a few ohms.

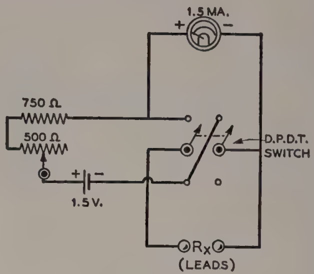


Figure 3.
Diagram of the low-range ohmmeter illustrated in Figure 2.

The calibration scale will depend upon the internal resistance of the particular make of 1.5-ma. meter used. The instrument can be calibrated by means of a Wheatstone bridge or a few resistors of known accuracy. The latter can be series-connected and parallel-connected to give sufficient calibration points. A hand-drawn scale can be pasted over the regular meter scale to give a direct reading in ohms.

Before calibrating the instrument or using it for measurement, the test prods should always be touched together and the zero adjuster set accurately.

Absorption-Type Wavemeter

The wavelength of any oscillator, doubler, or amplifier stage can be roughly determined with the aid of a simple absorption wavemeter. It is particularly useful for determining the correct harmonic from a harmonic crystal oscillator or frequency doubler or quadrupler. It consists of a simple tuned circuit which is

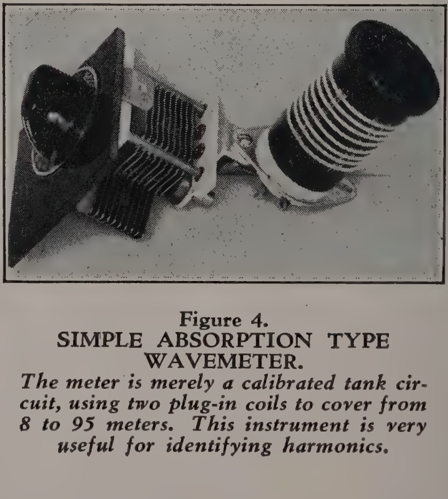


Figure 4.
SIMPLE ABSORPTION TYPE WAVEMETER.
The meter is merely a calibrated tank circuit, using two plug-in coils to cover from 8 to 95 meters. This instrument is very useful for identifying harmonics.

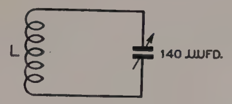


Figure 5.
ABSORPTION WAVEMETER CIRCUIT.
For the range of from 8 to 30 meters, L should consist of 8 turns 1" long on a 1¼" diameter form. For 30 to 95 meters, L should consist of 27 turns 1" long on a 1¼" diameter form.

coupled to the tank circuit under measurement. The wavemeter absorbs a small amount of energy from the transmitter tank circuit; this produces a change in reading of the milliammeter in the plate or grid circuit. A sharp rise or dip in the milliammeter current reading will take place when the wavemeter is tuned to the same wavelength or frequency as that of the circuit under measurement.

The coil socket is bolted to the back mounting flange of the 140-μμfd. midget variable condenser. One coil covers from 8 to 30 meters, another from 30 to 95 meters. The coil turns should be held in place with Duco cement.

The wavemeter can be calibrated by holding it near the secondary coil of an ordinary calibrated regenerative receiver which tunes to the known amateur bands. As the wavemeter condenser is rotated through its range, a point will be found where the receiver is pulled out of oscillation, as indicated by a sharp click in the headphones of the receiver. This point is then marked on the scale of the wavemeter dial. This calibration is sufficiently accurate to insure transmitter operation in the 10-meter band rather than 13-meter operation, which can be easily mistaken for 10-meter output when tuning a transmitter.

The wavemeter can also be calibrated by holding it near the plate coil of a crystal oscillator. A change in oscillator plate current or even a cessation of oscillation will occur when the wavemeter is tuned to the same frequency as that of the oscillator.

One can either make a continuous calibration curve for the two coils or make notes of the dial settings for the various amateur bands.

Should a u.h.f. range be desired, the 56- and 112-Mc. bands can be covered with a 2-turn coil. However, Lecher wires are ordinarily used for measurement at these frequencies.

112-Mc. Bandspread Wavemeter

This wavemeter operates on the same principle as the preceding one, but it is specifically designed for the 112-Mc. amateur band. The

Figure 6.
112-MC. ABSORPTION TYPE WAVEMETER.
This meter is mounted inside a Mason jar to preserve the accuracy of calibration. There is no danger of getting "bit" when the meter is brought near a high power tank circuit.

mechanical and electrical construction is shown in Figure 8. C_2 is a small trimmer condenser used as a band setter, and has a negative temperature coefficient. One of the new zero coefficient condensers should be even better.

The entire unit is enclosed in a glass jar having a screw type cover. The photographs and drawings clearly illustrate the method of assembly. The two upright Lucite bushings are tapped at each end for a 6-32 thread and mounted upon a support also of Lucite. The upper support fitting the inside of the jar cover may be fashioned from a piece of Presdwood.

A standard 3-plate, midget variable condenser is dismantled and rebuilt to have but 2 plates, the stator plate being adjustable. This is accomplished by replacing the original stator rods with 6-32 machine screws. When the correct spacing has been found, the single

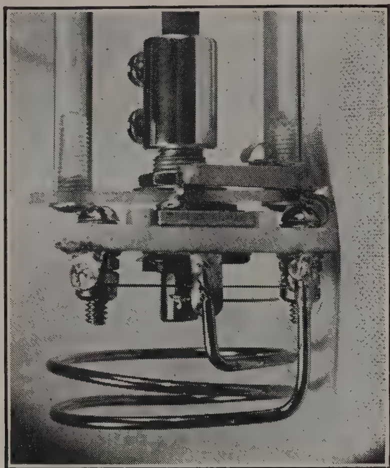


Figure 7.
CLOSEUP VIEW OF THE WAVEMETER.
This view clearly shows the construction. Note the Lucite posts.

stator plate is held rigidly in place by small hexagonal nuts.

A rubber gasket inside the jar cover makes the unit airtight. The two 6-32 machine screws extending through the dial plate also serve to hold the cover, rubber gasket, Presdwood support, and the upright bushings. The panel bearing is mounted in the exact center of the jar cover.

The pickup coil (L_1) consists of 2 turns of no. 12 wire, $1\frac{3}{4}$ inches in diameter and spaced approximately $\frac{1}{2}$ inch between the turns.

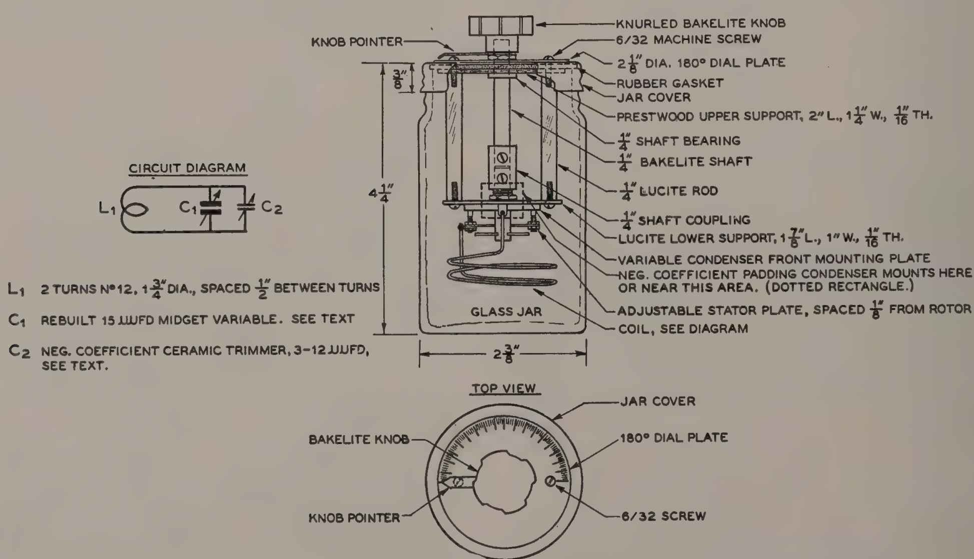
It is of paramount importance when assembling to make certain that all leads are short and rigid and that the rotor bearing of C_1 is making perfect contact.

Perhaps the most practical means of calibrating the instrument is by a Lecher wire measuring system coupled to a 112-Mc. receiver or transmitter, unless a pre-calibrated piece of equipment is already at hand. In either case, at least three readings should be taken identifying the center-, low-, and high-frequency limits of the 112-Mc. band. Other points may readily be found by interpolation.

If, during calibration, it is found that more bandspread will be required, the distance between the stator and rotor plate of C_1 may be increased. Ordinarily, this spacing will be approximately $\frac{1}{8}$ inch. This adjustment enables the entire 112-Mc. channel to be spread over a 180-degree arc or any position thereof.

It should be borne in mind that the accuracy of any absorption type wavemeter is limited by the fact that reactance is coupled into the wavemeter when it is brought near another tuned circuit, and also that the resonance peak covers quite a few kilocycles on the flat por-

Figure 8.
MECHANICAL AND ELECTRICAL CONSTRUCTION OF THE WAVEMETER.



tion of the "nose." For this reason, frequency measuring equipment of the heterodyne oscillator type should be used when a high degree of accuracy is required, the absorption meter being used simply to identify harmonics. In the latter case, the absorption wavemeter can be designed so that the band covers only a small portion of the dial. Then, when a new oscillator is being tuned up for the first time, it is possible not only to tell if it is out of the band, but in which direction and approximately how much.

Band Edge Frequency Spotter

The instrument diagrammed in Figure 11 and pictured in Figures 9 and 10 consists essentially of a 50-kilocycle oscillator and a tuned harmonic amplifier fed from a voltage-regulated power supply. It provides 50-kc. points of usable and adjustable strength on all the amateur bands up to and including 30 Mc. For that matter, it also gives usable calibration points every 50 kc. on the frequencies in between the amateur bands, should they be needed for some special purpose.

The 50-Kilocycle Oscillator. A 6K8 tube is used as a combined 50-kilocycle oscillator and electron-coupled doubler to 100 kc.

The oscillator coil is a Meissner 456-kilocycle beat-oscillator unit with the mica trimmer removed. By loading this comparatively high-frequency coil to the low frequency of 50 kc., the oscillator tuned circuit becomes quite high C. The stability with respect to tube and circuit temperature variations and plate voltage variations is greatly improved by the high oscillator lumped capacity. Actually, the capacity required to tune this oscillator coil to 50 kilocycles is very close to 0.00625 microfarads. This capacity is obtained through the use of a .006-μfd. fixed condenser of the silver-plated mica type in parallel with a .0002-μfd. negative coefficient condenser of the ceramic type and a 100-$\mu\mu$fd. midget variable. It is important that the identical coil as shown in the *Buyer's Guide* be used (and that the mica trimmer thereon be removed) if the values of capacity shown are to hit 50 kilocycles. It is also important that the specified type fixed condensers be used for lump capacity.

The 100-$\mu\mu$fd. midget trimmer condenser is brought through the chassis and allows the frequency of the oscillator to be tuned about one-half kilocycle either side of the operating frequency of 50 kilocycles. This adjustment will ordinarily compensate for any small variations in coil inductances and in circuit capacity. If, however, it is impossible to tune this circuit to 50 kc. by a variation in the capacity

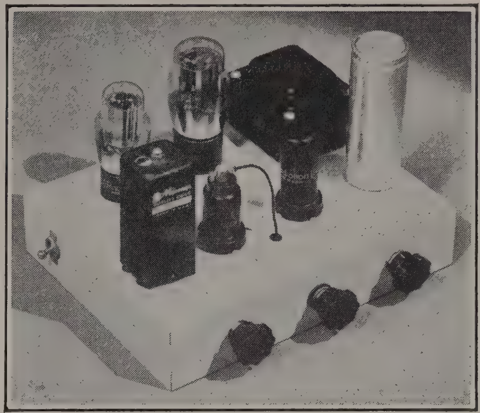


Figure 9.
FRONT VIEW OF THE FREQUENCY SPOTTER.
The control to the left of the front panel is the trimmer on the 50-kilocycle oscillator. The right-hand control is the harmonic amplifier coil switch, and the center control is the trimmer condenser across this coil.

of this condenser, the addition or subtraction of .00005 μfd. from the fixed value of .00625 will usually allow it to be accomplished.

The oscillator coil, as it comes from the manufacturer, is not a single tapped unit but rather a 2-winding affair with four leads brought out from the coils. It will then be necessary to series these two coils as shown in the manufacturer's diagram as the connection for an electron-coupled oscillator. The cathode of the oscillator-mixer tube is then connected to the tap. The actual connections to the coil are as follows: blue wire, ground; green wire, grid; red and black connect together and go to cathode.

The Harmonic Amplifier. The output circuit of this stage consists of a tapped coil which is resonated by a 100-$\mu\mu$fd. midget variable. With the whole coil in the circuit the output can be peaked at any frequency from about 7500 kc. down to about 3500 kc. Nevertheless, there is ample output with this coil in the circuit down through the broadcast band. It is not necessary to resonate the output circuit at these low frequencies; there is more than ample output for all measurements and for calibration.

With the switch in the second position, all but 9 turns of the inductance are shorted and the coil will resonate at any frequency from about 7000 kc. down through 18,000 kc. This tap peaks in the middle of the dial for strong signals on the 14-Mc. band. With the switch

on the last tap, with only 4 turns in the circuit, the circuit peaks up in the 28-Mc. band and for a considerable distance either side of it.

Coupling of the output circuit to the external load is accomplished by means of a .000025-μfd. mica condenser which connects between the plate of the 6V6 and the output terminal. The decrease in the reactance of this condenser with increasing frequency tends to equalize the signal strength output of the unit over a wide range of frequency.

A simple resistance-capacity filtered power supply using an 80 rectifier is used for plate voltage to the unit. Ample filtering for the harmonic amplifier stage is attained through the use of the RC filter. The VR-150-30 voltage regulator with its associated resistors and condensers supplies very pure direct current to the 6K8 oscillator and first multiplier.

Tuning Up and Calibration. If the oscillator coil specified has been used, and if the exact values of capacity specified have been placed across the coil, it is only necessary to get the oscillator going on the proper frequency of 50 kilocycles; when this is once done, all other adjustments become very simple.

For tuning up the frequency spotter, the only additional piece of equipment required is a calibrated broadcast receiver and a few incoming broadcast signals on frequencies that are integral multiples of 50 kc. With the oscillator operating (with the output coil on the no. 1 tap—all the coil in the circuit) run a wire from the output terminal of the spotter to the antenna post on the b.c.l. set and connect a small external antenna to the receiver.

With the trimmer condenser in the oscillator set to about mid-scale, tune the b.c.l. set to the low-frequency end of the dial and pick up the first harmonic of the oscillator that can be tuned in. Mark down the frequency of this harmonic as determined from the calibration of the b.c.l. set. Then tune to the next harmonic (they will be easy to identify because of their lack of modulation) and mark down its frequency as again determined from the b.c.l. set. Keep doing this until 8 or 10 points are determined. Then subtract each frequency from the next higher one all the way down the line and average the resulting differences in frequency between the harmonics. If any one of the differences falls very far out of line, recheck its frequency to see if an error has been made or to see if a harmonic has been missed.

If the average of all the differences in frequency falls very near to 50 kilocycles (say 48 to 52 kc.) the unit is ready for calibration. If not, the values of padding condenser across the oscillator coil will have to be changed.

When the oscillator has been adjusted very closely to 50 kc. by the "difference-between-harmonics" method, pick out a broadcast station that is operating on some multiple of 50 kc.; one in the vicinity of 550 to 1100 kc. will be the best. Tune in this station, turn on the oscillator, and adjust the beat between the harmonic of the oscillator and the broadcast station to zero. Then find another b.c. station on a harmonic of 50 kc. and see if it also is at

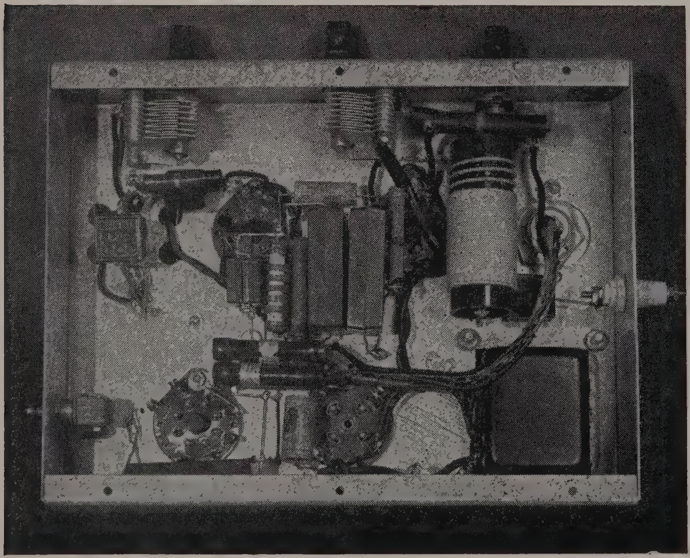
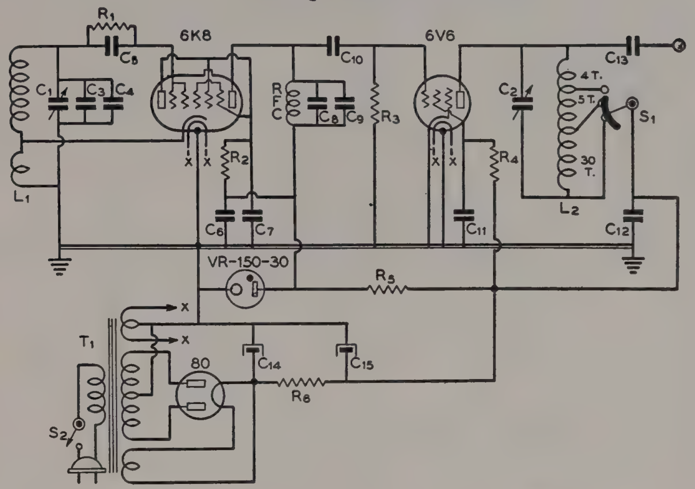


Figure 10.

BOTTOM VIEW OF THE CHASSIS SHOWING PLACEMENT OF COMPONENTS.

The tapped amplifier output coil is to the right and the r.f. choke with the two paralleled condensers across it, comprising the plate circuit of the 6K8, is just to the left of the center of the chassis.

Figure 11.
WIRING DIAGRAM OF THE FREQUENCY SPOTTER.



C_1, C_2—100-$\mu\mu$fd. midget variable
C_3—.006-μfd. silvered mica (close tolerance)
C_4—.0002-μfd. negative coefficient ceramic capacitor
C_5 — .0005 - μfd. midget mica
C_6, C_7—0.1-μfd. 400 volts
C_8—.001-μfd. mica
C_9—.0001-μfd. mica
C_{10}—.001-μfd. mica
C_{11}—0.1-μfd. 400 volts
C_{12}—.01-μfd. midget mica
C_{13}—25-$\mu\mu$fd. midget mica
C_{14}, C_{15} — 8-8-μfd. 450-volt elect.
R_1—100,000 ohms, 1 watt
R_2 — 25,000 ohms, 2 watts
R_3—2 megohms, 1 watt
R_4 — 50,000 ohms, 2 watts
R_5 — 20,000 ohms, 10 watts
R_6 — 10,000 ohms, 10 watts
S_1 — Shorting type 1-pole 3-position switch
S_2—S.p.s.t. a.c. line switch
L_1 — 456-kc. b.f.o. coil with trimmer removed
L_2 — 39 turns no. 20 d.c.c. on 1" dia. form tapped at 9 and 4 turns; last 4 turns spaced
R F C — 2.1 - mh., 1 2 5 - ma. r.f. choke
T_1—650 v. c.t., 40 ma.; 6.3 v. 2 a.; 5 v. 2 a.

zero beat. If the second b.c. station is not at zero beat or within a few cycles of it, the oscillator definitely isn't on 50 kc. and the frequency will have to be rechecked by the procedure given in the preceding paragraphs, fixed condensers being added or subtracted, depending on whether the frequency is high or low.

If the second station is at zero beat with the harmonic, check with a few more stations on multiples of 50 kc. just to make sure all is well. As mentioned before, if the values and components given are used, it will only be necessary to adjust the trimmer condenser across the oscillator tank, which is brought out to the front panel, to hit 50 kc. and thereby arrive at this stage of the adjustment.

It will now only be necessary to set the trimmer condenser so that the harmonics in the broadcast band fall exactly at zero beat with the b.c. stations, and the unit will thenceforward be calibrated.

It will be found that strong, steady signals are available every 50 kilocycles throughout all the amateur bands from 160 through 10 meters. These signals can be used as band-edge markers for either the 'phone or c.w. bands. Or, if a receiver with substantially straight-line-frequency tuning and an accurate dial is in use, the frequency of any incoming or locally-generated signal may be determined to a good degree of accuracy by interpolating between the 50-kilocycle points.

The warm-up time of the unit is very short, a matter of only 5 minutes or so, due to the very high value of capacity in the 50-kilocycle oscillator tank. Once the oscillator has been set, it will not drift more than a few cycles on the broadcast band (less than 100 cycles on 10 meters) in many hours of continuous operation. However, each time the unit is placed in operation from a cold start it will be best to check the setting of the oscillator trimmer condenser against a broadcast station on a multiple of 50 kc. after allowing 5 minutes or so to warm up.

Dual Frequency Crystal Calibrator

At a reasonable price, it is possible to purchase a crystal unit containing a crystal which will oscillate on either 100 or 1000 kc., oscillating along its length for the lower frequency and through its thickness for the higher frequency. The accuracy of the 100 kc. oscillation is very high, when installed and adjusted

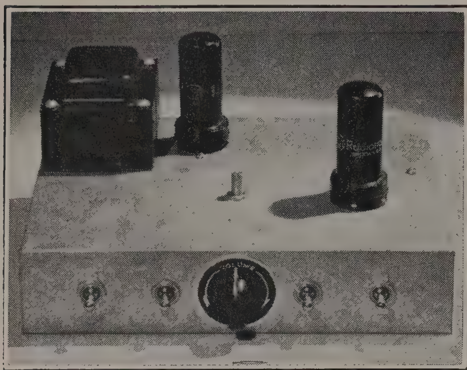


Figure 12.
DUAL FREQUENCY CRYSTAL CALIBRATOR.
This instrument generates 1000-kc. harmonics up to 56 Mc., and 100-kc. harmonics up to 20 Mc., the latter with a high degree of accuracy. On the front of the chassis is the output control; the slotted shaft projecting out of the top of the chassis is a shunt trimmer across the crystal for adjusting the low frequency oscillations precisely to 100 kc.

in a calibrator of the type to be described, permitting precision frequency measurement.

The advantage over the 50-kc. self-controlled unit just described is that it is much simpler to get going initially, and one need not allow for warm up or check it each time before taking a measurement. It has the further advantage that it can be made to oscillate with reasonable accuracy (.05 per cent) at 1000 kc., which is of considerable help in identifying the 100 kc. points on the higher frequencies where it is sometimes difficult to determine the order of a particular 100 kc. harmonic. The only disadvantage is that the 100 kc. points are twice as far apart as the points given by the "Band Edge Spotter" previously described, and the strength of harmonics above 14 Mc. is not quite as great with the crystal calibrator, due to the lack of a separate harmonic amplifier.

It should be borne in mind that the accuracy of the crystal when oscillating on 1000 kc. is not supposed to be sufficient for precision measurements; it is simply for convenience in identifying the highly accurate 100 kc. points.

Construction. The only precautions to be observed in construction are to place the crystal away from any components which radiate heat, and to make the leads from the tube to the crystal as short as possible. Mounting the crystal under the chassis close to the 6F6 socket, as illustrated, meets these requirements. The leads to condenser C_6 should also be kept as short as possible.

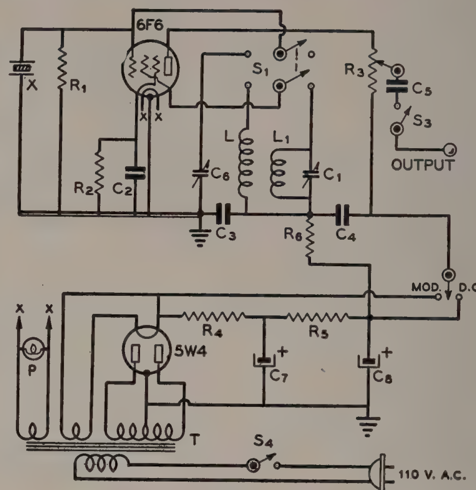


Figure 13.
WIRING DIAGRAM OF CRYSTAL CALIBRATOR.
(as recommended by Bliley Electric Co.)

X—100 and 1000 kc. crystal calibrator unit
R_1—5-meg., ½ watt
R_2—500 ohms, 1 watt
R_3 — 0.5-meg. potentiometer
R_4 — 10,000 ohms, 1 watt
R_5 — 20,000 ohms, 1 watt
C_1 — 25-100 $\mu\mu$fd. mica trimmer
C_2, C_3, C_4—0.1-μfd. tubular
C_5—.001-μfd. mica
C_6—25-$\mu\mu$fd. midget variable
C_7, C_8—Dual 4-μfd. 450-volt electrolytic
T — 320 v. each side of c.t., 40 ma. and fil. windings
S_1—D.p.d.t. toggle switch
S_2—S.p.d.t. toggle switch
S_3—S.p.s.t. toggle switch
L—R.F.C., exactly 8 mh.
L_1—Pie-wound 2.1 or 2.5 mh. 125 ma. r.f. choke with all sections except one removed

The coil L_1 is one pie of a midget 125 ma. 2.1 or 2.5 mh. r.f. choke. If the crystal will not oscillate with the switch on the 1000-kc. position, it may be necessary to remove a few turns from the coil. A midget solenoid b.c.l. coil may be substituted if desired.

Adjustments and Use. First check for oscillation on both positions of S_1 to make sure the crystal will oscillate both on 1000 and 100 kc. If it does not oscillate on 1000 kc., then the mica trimmer C_1 should be varied. When this trimmer has once been adjusted so that the crystal "comes on" every time the switch is thrown, the trimmer need never be touched again.

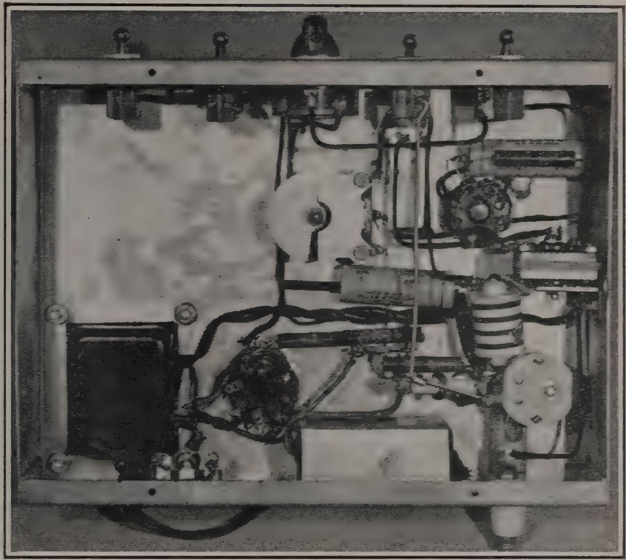


Figure 14.
UNDER-CHASSIS VIEW OF THE CRYSTAL CALIBRATOR.
The crystal unit is mounted close to the 6F6 sockets. Power supply components are to the lower left.

The 100 kc. frequency should then be precisely adjusted by means of the 25-$\mu\mu$fd. air trimmer, C_6, until harmonics of the 100-kc. oscillator zero beat *exactly* with broadcast stations on multiples of 100 kc. Zero beat can most accurately be determined if the receiver has a tuning eye or "R" meter. The trimmer is first adjusted as close as possible by ear and then further adjusted until the "flutter" of the indicator is reduced to as low a frequency as possible. The adjustment should preferably be checked against three or four broadcast stations and an average taken if there is any deviation for the different stations. No modulation should be applied to the oscillator during this adjustment.

The calibration when thus obtained will hold over a long period of time, and it is not necessary to check the frequency against broadcast stations before taking a measurement so long as the room temperature does not vary too much from the temperature at which the instrument was originally calibrated.

Modulation is accomplished by applying unfiltered r.a.c. to the output circuit of the oscillator, and can be cut in or out by means of the s.p.d.t. toggle switch indicated. The modulation facilitates spotting of the harmonics by making it easier to pick them out from among stray carriers.

Care should be taken with any 50-kc. or 100-kc. oscillator when making measurements on 14 Mc. or above if the receiver used does not have good image rejection. The appearance of images will result in spurious carriers and false readings.

Diode-Type Field-Strength Meter

The most practical method of tuning any antenna system, such as a half-wave antenna or a directional array, is by means of a field-strength meter. This instrument gives a direct indication of the actual field strength of a transmitted signal in the vicinity of the antenna. The device consists of a tuned circuit and a diode rectifier which is connected in series with a microammeter so that the meter will read the carrier signal strength.

A 0–200 microammeter as an indicator provides higher sensitivity than can be obtained with the more common 0–1 ma. meter ordinarily used for this purpose. A 0–100 microammeter will give still greater sensitivity.

The unit is inexpensive and requires but a single $1\frac{1}{2}$-volt cell for power. Besides serving as a field-strength meter, it can be used as a neutralizing indicator, or calibrated for use as an absorption wavemeter. The entire unit, except coils and coil socket, is housed in a metal can 6 inches cubical. The externally mounted coil facilitates coil changing and better adapts the unit for use as a wavemeter, no antenna or pickup wire being necessary in this application.

For service as a field-strength meter, the coil can be coupled to a small doublet by means of 2 or 3 turns of insulated wire wound around the coil. The instrument will be most sensitive if the pickup doublet is made resonant; but such a resonant doublet may, if it is closer than two or three wavelengths, upset the operation of the antenna being adjusted.

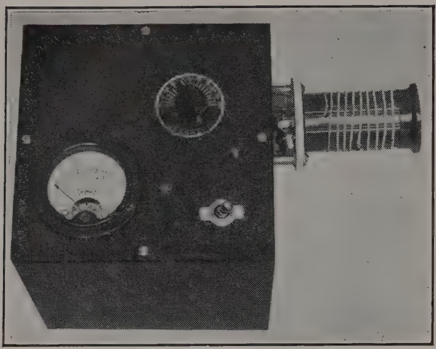


Figure 15.
SIMPLE FIELD STRENGTH METER.
This field strength meter uses a type 30 connected as a diode; a 0-100 or 0-200 microammeter provides good sensitivity. The unit can be used as an absorption wavemeter if calibrated. It also can be used as a neutralizing indicator.

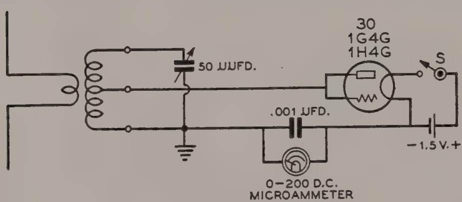


Figure 16.
WIRING DIAGRAM OF THE SIMPLE FIELD STRENGTH METER.

The instrument (Figure 15) was checked against a signal generator. With a type-30 tube, the following calibration in terms of decibels was obtained, using $12\frac{1}{2}$ μa. as an arbitrary zero db reference level:

$12\frac{1}{2}$ μa.— 0 db	100 μa.—15 db
25 μa.— 5 db	150 μa.—18 db
50 μa.—10 db	200 μa.—20 db

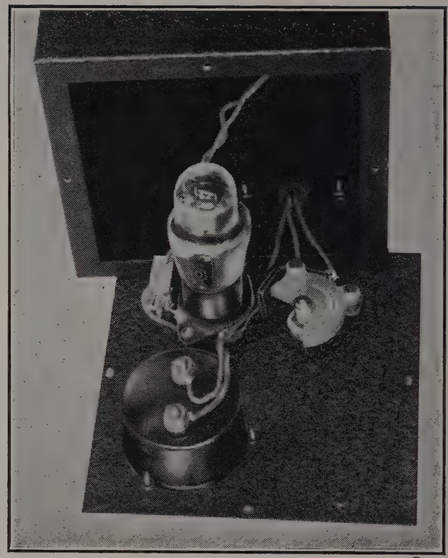


Figure 17.
INTERIOR VIEW.
A single "little six" dry cell is fastened inside the cabinet and supplies 1½ volts to the filament of the type 30 tube, used as a diode rectifier.

High-Sensitivity Field-Strength Meter and Simple V.T. Voltmeter

When it is desired to make field-strength readings some distance from the antenna, especially when a low-powered transmitter is used, the diode-type field meter just described does not have sufficient sensitivity. The field-strength meter illustrated is considerably more sensitive, but requires a plate battery and a more expensive tube than the diode type previously described.

A 1B4 tube, triode connected, is used as a detector. Two small batteries are required for

COIL TABLE

160 λ	88 turns # 26 d.c.c. 1½" diam. closewound center tap	20 λ	10 turns # 22 d.c.c. 1½" diam. 1½" long center tap
80 λ	38 turns # 22 d.c.c. 1½" diam. closewound center tap	10 λ	6 turns # 22 d.c.c. 1⅛" diam. 1" long center tap
40 λ	24 turns # 22 d.c.c. 1½" diam. 1½" long center tap	5 λ	2 turns # 18 enam. 1⅛" diam. ¾" long center tap

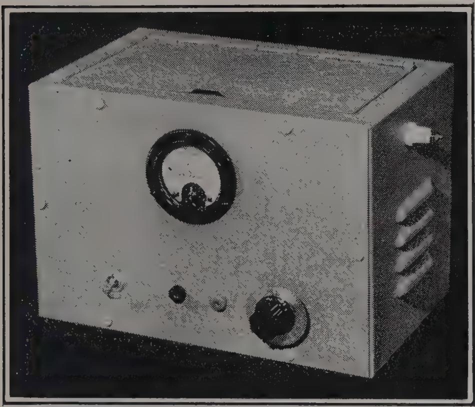


Figure 18.
SENSITIVE FIELD STRENGTH METER.
This is a more sensitive device than the one illustrated in Figures 15, 16, and 17, and also can be used as a simple vacuum tube voltmeter. It requires more batteries, however.

the plate, filament and bias supplies. The plate voltage is 22½ volts, the bias about 2½ volts, and the rated filament voltage, 2 volts.

The one tuned circuit in the meter is designed to cover any two consecutive amateur bands.

In the unit shown, one coil is used to cover 10 and 20, another to cover 40 and 80, and still another coil to cover the 160-meter band. All coils are wound on 1½-inch coil forms.

The 10–20 meter coil contains 4 turns spaced to about ¾ inch; the 40–80 meter coil is of 15 turns, close-wound; and the 160-meter coil is of 50 turns, close-wound. All coils are wound with no. 20 wire; the 10–20 one is wound with enamelled and the other two with d.c.c. wire. Both of the 2-band coils (10–20 and 40–80) will hit the lower-frequency band with the condenser plates almost completely meshed, and the higher-frequency band with them almost separated.

When the unit is first turned on, if the batteries and the tube are in good condition, the 0–1 d.c. milliammeter in the common plate and screen circuit of the 1B4 will indicate about 50 microamperes of plate current.

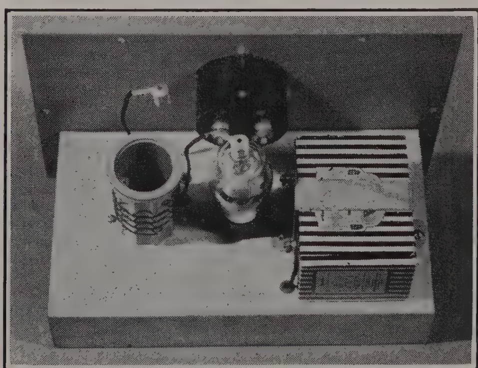
Now, to return the meter to the zero position, with this .05-ma. flow going through it, it is only necessary to turn the zero-adjustment screw until the needle points to zero with the meter in operation. Then, the fact that the meter will always point to zero when all components are in adjustment will serve as a check on the calibration and condition of the batteries. However, as soon as the meter is turned off, the pointer of the milliammeter will fall below the zero on the scale. (Actually it will rest upon the pin on the zero side of the meter.)

If a short length of wire is used for pickup, it may be connected directly to the antenna post, which is a stand-off insulator on the side of the cabinet, wired directly to the stator of the tuning condenser. If it is impossible to get a substantial deflection with a short length of wire, the case of the instrument should be grounded and a longer piece of wire connected to the antenna post through a 3–30 $\mu\mu$fd. mica trimmer used as a pickup. If the trimmer is not used, a long pickup antenna will detune the meter.

The tuning condenser C_1 can be detuned from resonance if too great a deflection is obtained. It is not necessary that the tank be tuned to resonance for field-strength measurements, though the meter will be most sensitive when C_1 is tuned to exact resonance.

For most work, calibration is not required, a relative indication being sufficient. However, the dial may be calibrated in decibels if desired. The decibel calibration may be marked directly on the meter scale, or a calibration chart may be made. To calibrate the instrument in decibels, simply reduce the input to a class C amplifier in given steps after adjusting the f.s. meter for full scale deflection. Cutting the plate voltage to the class C stage in half would be a power reduction of four times, or 6 db, and so on. The meter covers a useful range of approximately 20 db.

Figure 19.
REAR VIEW OF SENSITIVE FIELD STRENGTH METER.
The plate battery is mounted above the chassis, held firmly in position by means of a strap. Note the extra grid clip, which is brought out to a pin jack for v.t. voltmeter use.



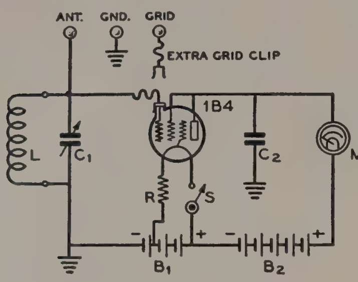


Figure 20.
WIRING DIAGRAM OF SENSITIVE FIELD STRENGTH METER.

C_1—140-$\mu\mu$fd. midget variable
C_2—.002-μfd. mica
R—15-ohm resistor
S—On-off switch, s.p.s.t. toggle
M—0-1 d.c. milliammeter
B_1 — 4½-volt C battery. Filament leads connected to + and — 3, ground connected to — 4½
B^2 — 22½-volt C battery
L—Coils—see text

If the instrument is used as a wavemeter, the dial calibration should be made with a short, rigid piece of wire as a pickup. The same wire should then be used whenever subsequent wavelength measurements are made. This wire or rod need not be over a few inches long, as it will receive sufficient pickup when brought near the tank circuit whose frequency is to be determined.

For simple vacuum tube voltmeter measurements it is only necessary to substitute the extra grid clip and make connections from the pin jacks to the device whose voltage is to be measured.

When used as a v.t. voltmeter the instrument should be calibrated by means of an adjustable a.c. voltage supply of 0–5 volts and a 5-volt a.c. voltmeter. If desired, the voltage calibration can be converted to decibels, thus making it unnecessary to calibrate it by the method previously described.

Grid-Leak F.S. Meter

Slightly greater sensitivity and decibel range can be obtained with a grid-leak type f.s. meter than with the power detector type just described. However, it requires a higher voltage plate battery and has the disadvantage that it cannot be used as an accurate v.t. voltmeter. But if the transmitter power is low, or if readings must be taken at considerable distance, the grid-leak type is to be preferred. It will give usable readings on approximately 10 db weaker signals than will the power detector type previously described.

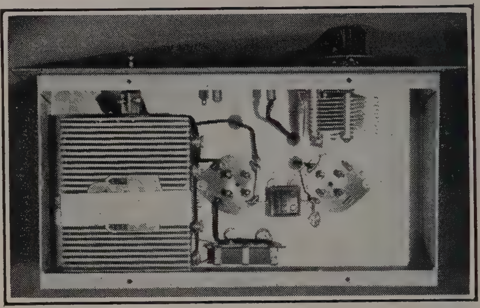


Figure 21.
UNDER-CHASSIS VIEW OF SENSITIVE F.S. METER.
The combined "A" and "C" battery is strapped under the chassis as shown here.

The instrument illustrated in Figure 22 and diagrammed in Figure 23 utilizes a type 19 dual triode for maximum power sensitivity, though any 1.5-volt or 2-volt triode having a μ between 20 and 50 will be satisfactory if the plate voltage is changed accordingly. The plate voltage should be just sufficient to make the tube draw very close to 1 ma. when no excitation is applied. The meter is then adjusted to exactly 1 ma. by means of the zero-adjuster screw. A signal will cause a *deflection* in plate current, the amount of *deflection* increasing with an increase in signal input voltage up to a saturation point corresponding to about 0.25 ma. scale. To make the meter pointer read forward with increased signal strength, the meter is mounted upside down.

The instrument illustrated can be used from 10 to 160 meters. For 5 and 2½ meters, an HY-114 should be substituted for the 19 or 1J6-G, and a 15-$\mu\mu$fd. sub-midget variable should be substituted for C_1. The mechanical arrangement should be changed to allow shorter tank circuit leads for 5 and 2½ meters. The required plate voltage for a no-signal plate current of 1 ma. will be about the same as for a 19, or about 50 volts.

The required voltage must be obtained by changing the plate battery voltage, adding in 1½-volt steps, if necessary. A variable dropping resistor in the plate voltage lead is not satisfactory, as the voltage drop through this resistor will decrease as the plate current decreases, thus reducing the sensitivity of the instrument.

A fairly accurate calibration may be obtained by using the accompanying calibration chart, provided a type 19 or 1J6-G is used. Assuming the meter has 50 scale divisions, start counting from zero signal or 1 ma. and mark the meter face in pencil to correspond

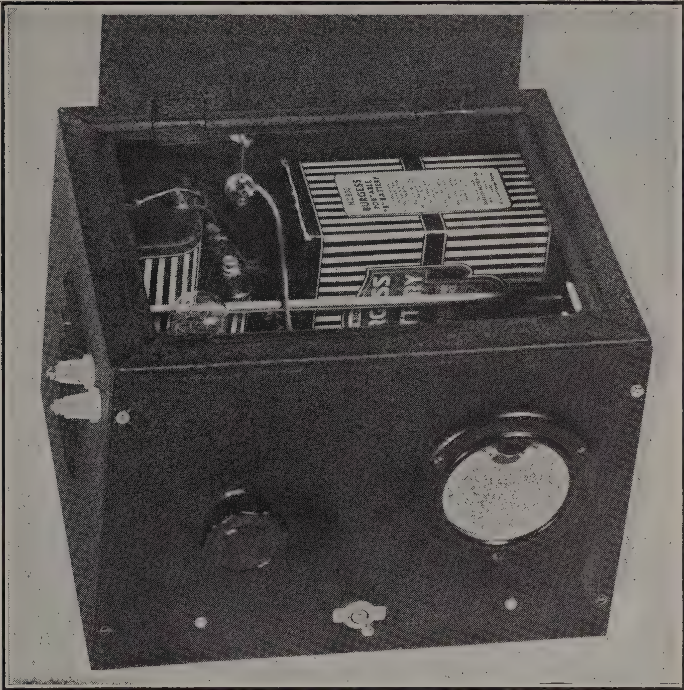


Figure 22.

HIGHLY SENSITIVE, GRID-LEAK TYPE F.S. METER.

The doublet feeder terminals are on the side of the cabinet, and the antenna terminal is on the rear of the cabinet to reduce body capacity. Note the inverted method of mounting the 0-1 milliammeter and hand calibrated db scale.

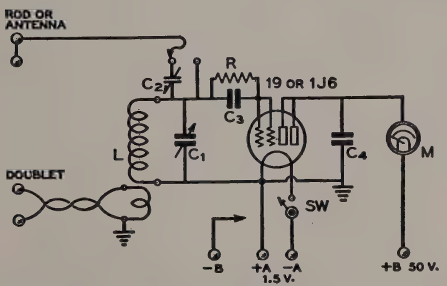


Figure 23.

GENERAL WIRING DIAGRAM OF THE F.S. METER.

The B minus is connected to that side of the A battery which gives a plate current reading closest to 1 ma. with the plate meter inverted. Be sure A positive is connected to chassis (ground).

L—See coil table
C_1 — 50-μμfd. midget condenser for 5-40 meters (25-μμfd. for 2½-20 meters)
C_2 — 3-30-μμfd. mica trimmer
C_3—.0005-μfd. midget mica condenser
C_4—.001-μμfd. midget mica condenser
M — 0-1 ma. 3 in. round (bakelite) case milliammeter
R — Accurate 5-meg-ohm ½-watt resistor

to the calibration chart. (With 50 scale divisions, 1 division is equal to 20 μa. or .02 ma.)

If greater accuracy is required, the meter should be calibrated individually as described for the meter of Figure 18.

Care should be taken with the 0–1 ma. meter when installing it to make sure the negative meter terminal isn't accidentally shorted to ground (by the pliers or screwdriver touching the cabinet, or other means). The meter will be blown instantly by the B voltage if this happens. It is best to disconnect the B plus lead to the plate battery when connecting or disconnecting leads to the meter. A small instrument-type fuse in series with the meter is desirable for protection.

Scale Divisions	Db Calibration
3	0 db
5	3 db
8	6 db
12	9 db
17	12 db
22	15 db
27	18 db
31	21 db
34	24 db
36	27 db

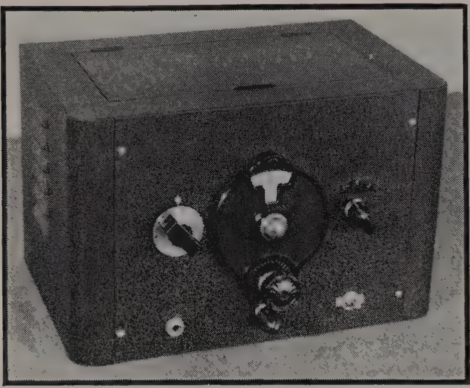


Figure 24.
FREQUENCY METER, MONITOR FOR C.W., AND EMERGENCY C.W. RECEIVER.
This unit is completely self-contained and self-powered. As a heterodyne frequency meter it is sufficiently accurate for amateur work. Bandswitching is used to eliminate the need for changing coils.

It should be borne in mind that this f.s. meter has a "saturation point," beyond which the deflection flattens off and then actually decreases with further increase in signal strength. In taking a reading, care should be taken to make sure the signal is not so strong as to block the meter, as a deceptive indication then will be obtained. Detuning the tank condenser from resonance should always result in less signal strength; if this does not occur, the signal is blocking the f.s. meter and it must be cut down before a reading can be taken.

Monitor, Frequency Meter, and Emergency C.W. Receiver

A c.w. monitor is a useful adjunct to a c.w. station as a means of checking the emitted signal for chirps, excessive ripple, key clicks, tails, and other undesirable characteristics. A shielded monitor enables the operator to tell from within the station what the signal sounds like at a distance. The simplest c.w. monitor is a modified autodyne receiver, well shielded.

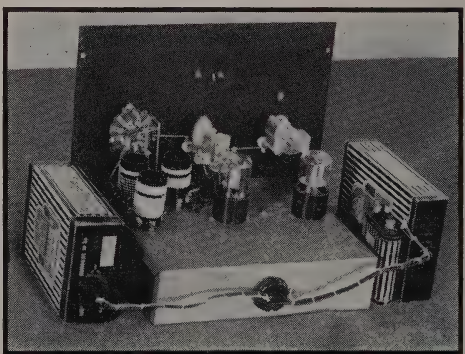
If the device is stable and well constructed mechanically, it can also serve as a heterodyne frequency meter. An accurate means for determining the frequency of a radio transmitter is essential when the circuit is of the self-excited oscillator type, and useful when the transmitter is of the crystal controlled type, to make certain that the crystal is not oscillating on a spurious frequency. In fact, the government regulations require that amateurs have some such means of measuring their frequency.

The multi-purpose unit shown in Figure 24 may be used either as an accurate heterodyne frequency meter, as a c.w. monitor, or as an emergency receiver suitable for c.w. reception. It covers the 80-, 40-, and 20-meter bands by means of coil switching. Battery powered, the unit is entirely self-contained, the necessary batteries being housed inside the cabinet. As the filament drain is only 0.1 amp. and the total plate current drain about 2 ma., the batteries will last a long time. The output is sufficient to drive an efficient loudspeaker to moderate room volume, but the unit is designed primarily for use with earphones (magnetic type). A closed circuit earphone jack is used so that the tube will not be running with screen voltage but without plate voltage when the earphones are removed, thus preventing damage to the tube.

A rather low value of detector grid-leak is used, and this reduces the sensitivity slightly. However, this is necessary in order to prevent the tube from blocking too easily when the device is employed as a keying monitor. The sensitivity is still sufficient to permit reception of moderately weak c.w. signals when the unit is used as an emergency receiver.

It will be noted that no regeneration control is employed. Such a control would affect the calibration, and therefore would be undesirable when the set was used as a frequency meter. Rather than include one, with provision for cutting it out when the unit is used for frequency measurement, the regeneration is

Figure 25.
FREQUENCY METER-MONITOR, INTERIOR CONSTRUCTION.
The cabinet is chosen large enough to accommodate the various batteries. The A battery is good for 300 hours' actual operation, the B and C batteries much more.



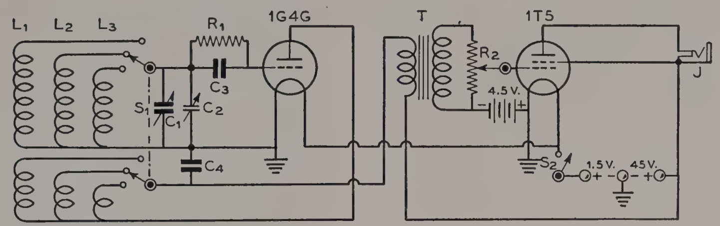


Figure 26.
WIRING DIAGRAM OF MONITOR-FREQUENCY METER UNIT.

C_1—35-$\mu\mu$fd. midget condenser (main tuning)
C_2—140-$\mu\mu$fd. midget condenser (bandset)
C_3—100-$\mu\mu$fd. midget mica
C_4—.006-μfd. midget mica
R_1—250,000 ohms, ½ watt
R_2—500,000-ohm pot., a.f. taper (vol. control)
T—Midget or replacement type 1:3 ratio a.f. transformer
J—Closed circuit jack, insulated from panel
Coils—See coil table
S_1—Double-pole 3-position selector switch

fixed at an optimum compromise value. Because of its effect upon calibration, provision for antenna coupling likewise was omitted. Should the unit be required for emergency use as a portable battery powered receiver, an antenna is coupled directly to the grid by means of a small compression trimmer or by wrapping a few turns of insulated wire around the grid lead on the coil side of the grid condenser.

With the switch on the 80-meter position the unit may be used as a monitor for either 80- or 160-meter signals. With the switch on the 20-meter position, the unit may be used as a monitor for either 20- or 10-meter signals. Sufficient pickup is obtained even though the detector is tuned to an adjacent band. For frequency meter use, the unit *always is operated on 80 meters.* Harmonic operation permits frequency measurement on other bands. The condensers are of such capacity that the 3500- to 4100-kc. range is covered in two jumps (two settings of the bandset condenser). This spreads out the frequency divisions and makes closer readings possible.

The dial should be of the vernier type, of good quality and capable of being read to a small fraction of a division by interpolation. This permits readings accurate to better than 1 kc. at 80 meters, which is as close as the meter should be depended upon anyway. There is no point in being able to read a frequency meter to 100 cycles if the accuracy of the meter itself cannot be depended upon to closer than 500 cycles.

The unit is calibrated with the aid of broadcast station harmonics, crystals of known frequency, or harmonics of a 100 kc. crystal standard. As many points as possible are obtained and two curves drawn on graph paper, one curve covering from 3500 to about 3775 kc. and the other curve from 3775 to 4100 kc.

COIL TABLE FOR FREQUENCY METER-MONITOR

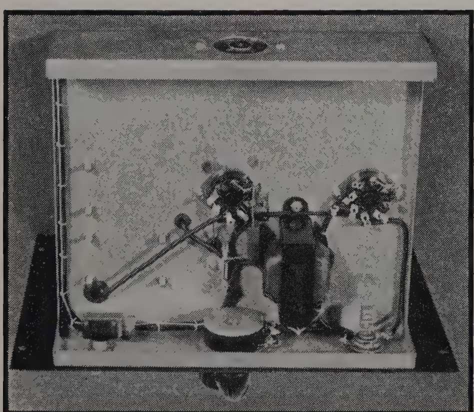
All coils are wound on 1 in. dia. bakelite forms 1¾ in. long, mounted to chassis by spade bolts. All coils wound with no. 24 d.c.c. Tickler polarity must be correct for oscillation.

L_1—80-meter coil. 23 turns close-wound. Tickler 12 turns close-wound, spaced ⅛ in. from grid coil.

L_2—40-meter coil. 12 turns close-wound. Tickler 5 turns close-wound, spaced ⅛ inch from grid coil.

L_3—20-meter coil. 8 turns spaced to ¾ in. Tickler 4 turns close-wound, spaced ⅛ in. from grid coil.

Figure 27.
UNDER-CHASSIS VIEW OF FREQUENCY METER MONITOR.



Both curves may be plotted on a single graph.

Because it is impossible to return *exactly* to the same bandset condenser capacity simply by turning the knob to the same position, and as various components tend to age and cause a slight change in frequency, some sort of *reference* signal is necessary for setting the bandset condenser each time the meter is to be used. This may be a low-drift crystal, a broadcast harmonic, etc. The measuring condenser, C_1, is set at the reading indicated for that frequency by the calibration graph, then the bandset condenser is slowly tuned near the known approximate setting until the signal is exactly zero beat. The meter then is ready for use. No "warming up" period need be observed before using the meter after it is turned on, as the tubes dissipate so little heat that warm-up drift is infinitesimal.

When measuring one's own transmitter frequency, it is permissible either to listen to the output of the device directly, as is done for monitoring, or it can be used simply as a heterodyne oscillator and the beat note observed in the station receiver. If a sufficiently strong beat is not obtained, the lid to the cabinet may be raised, in which case the calibration curve should be set again against the "reference" signal by means of the bandset condenser. Raising the lid changes the oscillator frequency slightly.

Mechanical construction is of considerable importance; all parts and wiring should be rigid and firmly anchored. As the battery type tubes are somewhat microphonic, it is recommended that the unit be placed on a sponge rubber pad.

The batteries should be wedged firmly in place with small pieces of wood, as the calibration will be affected if they are permitted to rattle around. A calibration oscillator suitable for obtaining calibration plotting points is shown in Figure 28.

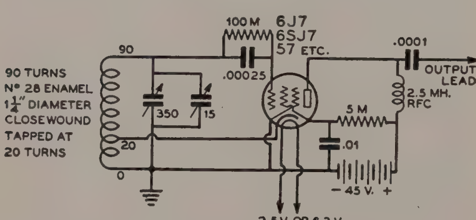


Figure 28.

CALIBRATION OSCILLATOR FOR A.C. TUBES.

With heater type tubes, battery plate supply is still used with the calibration oscillator for reasons of stability.

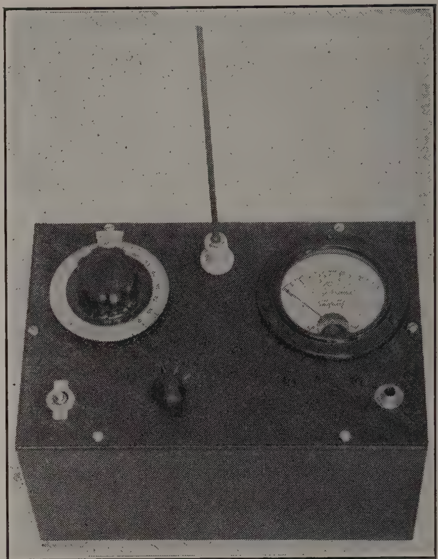


Figure 29.

'PHONE TEST SET.

This versatile instrument can be used as a phone monitor, an absorption wavemeter, field strength meter, overmodulation (carrier shift) indicator, or neutralizing indicator.

General Purpose 'Phone Test Set

A 'phone test set is quite similar to a field-strength meter, yet it lends itself to making additional measurements. It can be used as an overmodulation indicator, 'phone monitor, field-strength meter, neutralizing indicator, and wavemeter.

Such an instrument enables the operator to check for overmodulation of a 'phone transmitter. When the tuned circuit of the test set is coupled to the modulated amplifier or antenna system in such a manner as to obtain half-scale deflection of the milliammeter, any flicker of the meter reading will then be an indication of overmodulation. A change in meter reading during modulation is an indication of *carrier shift;* often this is a result of incorrect operating conditions, which may produce illegal interference in adjacent radiophone channels.

The 'phone test set consists of a diode rectifier connected across a tuned circuit, as shown in Figure 31. A 0–1 d.c. milliammeter serves to check overmodulation, and is useful as an indicator in field-strength measurements or neutralizing adjustments in a transmitter.

The audio volume with half to full scale meter indication is sufficient to give normal

headphone response. A 5,000-ohm resistor is connected into the jack circuit for use when the test set functions as an overmodulation indicator. This resistor is in series with the diode and tends to produce a more linear rectification of the carrier wave.

For neutralizing or field-strength measurements, a short-circuiting plug or brass rod should be inserted into the phone jack to short-circuit the 5,000-ohm resistor and thereby increase the sensitivity of the meter. Neutralizing adjustments are made by coupling the test set's tuned circuit to the transmitter stage under test (without plate voltage applied to the stage). When the stage is completely neutralized, there will be either a minimum or zero deflection of the meter needle.

A short piece of brass rod, about 10 inches long, protrudes from the chassis as may be seen in Figure 29; this rod acts as a pickup. For most purposes the signal pickup with this rod will be sufficient, but when the instrument is used for measuring field strength and there is insufficient meter deflection for an accurate reading, an auxiliary antenna consisting of several feet of insulated wire may be coupled to the pickup rod by wrapping one end of the insulated wire around the pickup rod a few times. The small amount of capacity coupling provided will be sufficient to give a higher meter reading, but will not be enough to disturb the frequency of the tank circuit appreciably.

When using the instrument in the neutralization of an r.f. amplifier, a short piece of flexible wire, about 18 inches long, is clipped directly to the pickup rod. The other end of the wire is brought closer and closer to the plate lead of the stage being neutralized until a substantial deflection is obtained.

Coil Data. The use of a 140-$\mu\mu$fd. tuning condenser permits use of one coil for 5 and 10 meters, another for 20 and 40 meters, and another for 80 and 160 meters.

For 5 and 10 meters the coil consists of 5 turns of no. 14 wire, $\frac{1}{2}$ inch in diameter and spaced to occupy a length of 1 inch. This coil is self-supporting and is soldered directly to the coil switch and tuning condenser rotor.

The 20–40 meter coil consists of 14 turns of no. 22 d.c.c. wire spaced to 1 inch on a $1\frac{1}{8}$-inch diameter form.

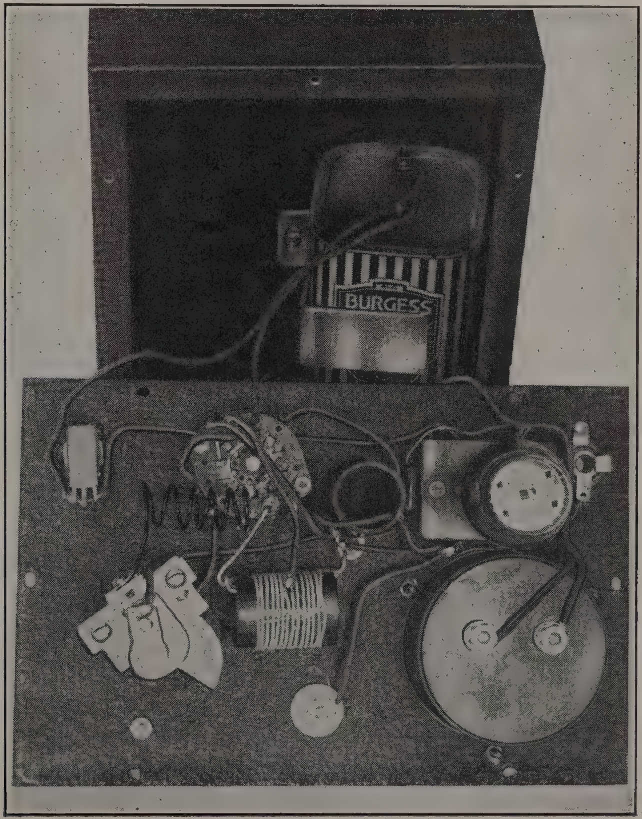


Figure 30.
INTERIOR CONSTRUCTION OF 'PHONE TEST SET.
All components are supported by the front panel; the single dry cell used for filament supply is fastened to the cabinet as illustrated.

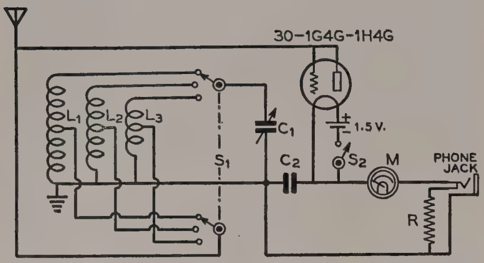


Figure 31.
WIRING DIAGRAM OF THE 'PHONE TEST SET.

C_1 — 140-$\mu\mu$fd. midget	S_1—2-pole 3-position rotary switch
C_2—.001-μfd. mica	S_2—Toggle switch
L_1, L_2, L_3—See text	M—0-1 ma. d.c. 3" meter
R—5000 ohms, ½ watt	

The 80–160 meter coil has 55 turns of no. 26 enamelled wire, close-wound on a $1\frac{1}{8}$-inch diameter form.

Calibration. If the instrument illustrated is duplicated carefully, there will be no need for plotting a calibration curve or table for the individual meter in terms of decibels. The following table will be sufficiently accurate (arbitrary zero db reference level taken as .05 ma. deflection).

0.05 ma.— 0 db	0.60 ma.—16 db
0.10 ma.— $4\frac{1}{2}$ db	0.70 ma.—17 db
0.20 ma.— $8\frac{1}{2}$ db	0.80 ma.—18 db
0.30 ma.—11 db	0.90 ma.—19 db
0.40 ma.—13 db	1.00 ma.—20 db
0.50 ma.—$14\frac{1}{2}$ db	

An individual frequency calibration must be made to cover use of the instrument as an absorption wavemeter. As a wavemeter, the instrument should be used only for rough measurements, such as determining the order of a harmonic.

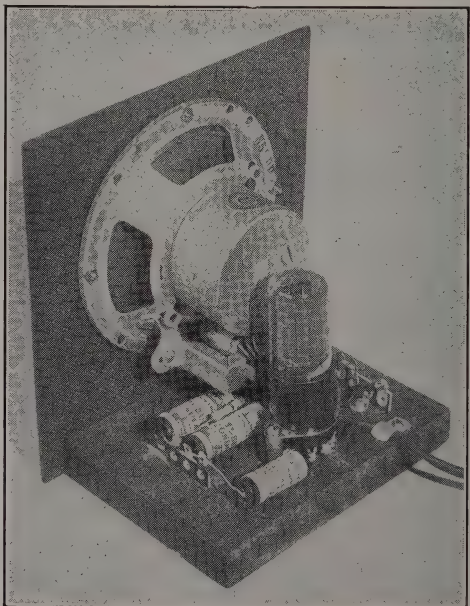
Keying Monitor and Code Practice Oscillator

The simple device illustrated in Figure 32 has many uses. It may be used as a keying monitor, facilitating accurate sending of the code characters (especially useful with a "bug" key) and as a "watchdog" on the character of the note. Any ripple or keying chirp present in the carrier in sufficient degree as to be objectionable is readily apparent on the monitor. It also may be used as a code practice oscillator. The speaker itself, requiring no external field supply, will come in handy around the test bench for use as a test speaker. To give the device this wide utility, several connections are brought out to terminals.

The speaker is a 5-inch p.m. dynamic type, complete with midget push-pull output transformer. The output transformer acts as the oscillation transformer for the tetrode section of the 117L7GT, which is used as a conventional Hartley oscillator. For plate voltage, some r.f. carrier voltage is picked up from the final amplifier plate coil by a few turns of heavily insulated wire and fed to the monitor by means of a twisted-pair or coaxial line. This carrier is rectified by the rectifier section of the 117L7GT and utilized as plate voltage. The plate by-pass condenser C_3 filters the rectified carrier into pure d.c. if the carrier itself is free from ripple. However, the time constant of this condenser is fast enough that any ripple in the carrier will show up as modulation of the signal generated by the keying monitor. Likewise, any keying lag will be apparent, because the strength of the monitor signal is determined by the strength of the carrier.

The amount of r.f. picked up by the pickup coil is adjusted until the monitor signal is of the desired volume. The r.f. power required is small, less than 1 watt for full room volume. For keying monitor use, the terminals marked "external B bat." should *not* be shorted. With

Figure 32.
KEYING MONITOR AND CODE PRACTICE OSCILLATOR.
This versatile unit may be used as a c.w. monitor, an audio "howler" for code practice, or as a test speaker requiring no field supply.



them shorted, damage to the tube may result.

For use as a code practice oscillator, a small B or C battery is connected in series with a key. A 22½-volt battery will give good room volume, and a fair signal is obtained with as little as 4½ volts. The current drain is low and the battery will have long life.

The tone or pitch of the oscillator can be varied by changing the value of C_1. A smaller capacity will give a higher pitched note, and

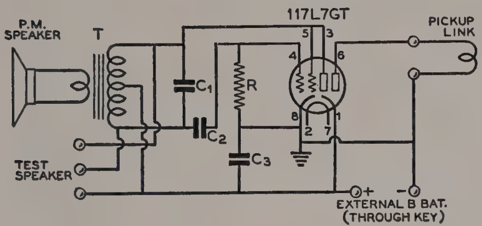


Figure 33.
WIRING DIAGRAM OF VERSATILE KEYING MONITOR.
The line voltage is applied directly to the heater of the 117L7GT, prongs 2 and 7 on the octal socket.

T — Midget push-pull output transformer (on speaker)
C_1—.01-μfd. (smaller capacity will give higher pitched note)
C_2—.05-μfd. tubular
C_3—0.1-μfd. tubular
R — 25,000 ohms, ½ watt

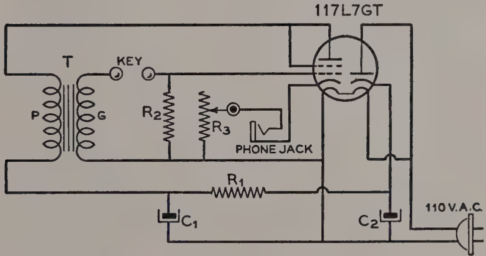


Figure 34.
A.C. OPERATED AUDIO OSCILLATOR.
For code practice only, this unit is more convenient, since no batteries are required. If the circuit fails to oscillate, the leads to one side of the audio transformer should be reversed. A high-pitched oscillation when the key is up may be stopped by changing the value of R_2.

C_1, C_2 — 10-μfd. 250-volt electrolytic
R_1—400 ohms, 10 watts
R_2—500,000 ohms, 1 watt
R_3—2-megohm potentiometer
T — Small single-plate - to - grid transformer

vice versa. If the condenser is made too large, however, the tube will no longer oscillate.

A "loose" speaker requiring no field supply is often useful for test purposes. By bringing out leads from the three primary wires as shown in the diagram, the speaker may be used for such purposes. For such work, the heater of the 117L7GT is not lit.

When used as a test speaker the highs will be somewhat attenuated in the manner of "tone control" because of the effect of the shunt condenser C_1. If suppression of the extremely high voice frequencies is undesirable, provision for opening one lead to C_1 can be made.

If the speaker transformer is of the variable ratio type, the voice coil tap should be chosen to give 14,000 ohms across the full primary, though this adjustment is not especially critical. More volume can be obtained for a given plate voltage by adjusting the voice coil tap for a lower primary impedance, but if this is carried too far the tube will not oscillate at low plate voltage. To give a true replica of the monitored signal, the monitor should be capable of oscillating on as little as 3 volts.

The unit is constructed on a small wooden baseboard and a Masonite front panel. It may be enclosed in a small cabinet or wooden box if desired.

A.F. and R.F. Power Measuring Device

For accurate measurement of a.f. and r.f. power, a thermogalvanometer in series with a non-inductive resistor of known resistance can be used. The meter should have good accuracy, and the exact value of resistance should be known with accuracy. Suitable dummy load resistors of the "vacuum" type are available in various resistances in both 100- and 250-watt ratings. These are virtually non-inductive, and may be considered as a pure resistance except at ultra-high frequencies. The resistance of these units is substantially constant for all values of current up to the maximum dissipation rating, but where extreme accuracy is required, a correction chart of the dissipation coefficient of resistance (supplied by the manufacturer) may be employed. This chart shows the exact resistance for different values of current through the resistor.

If a current-squared r.f. galvanometer (commonly 115-ma. full-scale deflection reading 0–100) is used, it will be necessary to use a conversion chart to determine the exact value of current from the scale reading. If a thermo-ammeter is used, the current reading can be taken directly from the meter.

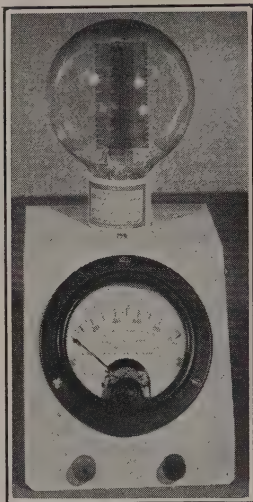


Figure 35.
R.F. AND A.F. POWER MEASURING DEVICE.
A thermogalvanometer or thermoammeter is placed in series with a non-inductive dummy load resistor whose resistance is known accurately.

Bandswitching Signal Generator

A signal generator is a useful piece of apparatus for aligning a receiver or calibrating a frequency meter from broadcast harmonics. When the generator is used for calibration purposes, the oscillator must be quite stable if a high degree of accuracy is required. In alignment work, a modulated signal often is desirable.

The signal generator illustrated in Figure 36 and diagrammed in Figure 37 meets these specifications. A fixed padder condenser (low temperature coefficient type) provides a minimum value of tank capacity which permits high stability at all tuning dial settings. A voltage regulator tube holds the plate voltage virtually constant in spite of line voltage fluctuations; hence the carrier is very stable after a short "warm up" period.

For a modulated signal, as sometimes is desirable for initial or course alignment, the filter and voltage regulator (which also provides filtering) are disconnected by means of S_2. This applies a half-wave rectified voltage which is rich in harmonics, the latter being sufficiently high in frequency to permit their audibility in receivers having poor bass response, yet low enough in frequency that the modulated signal is quite sharp. Because of the inherent stability of the oscillator, very little "wobbulation" takes place as a result of the modulated plate voltage.

The layout illustrated need not be adhered to closely; the only requirement is that r.f. leads be rigid and that the coils, tuning condenser, etc. be mounted firmly.

The range from 440 kc. to 1500 kc. is covered in two parts, the dividing line being at about 750 kc. To facilitate the obtaining of exact zero beat, a small vernier condenser is connected across the main tuning condenser. This condenser has only one rotor and one stator plate, double spaced. Almost any midget condenser may be adapted by removing plates as required.

Coils. The two coils are wound on 2-inch diameter bakelite tubing with no. 24 enamelled wire. The high frequency coil has a plate winding of 36 turns, close-wound, and a tickler winding of approximately 35 turns, close-wound. The two windings are separated about $\frac{1}{8}$ inch.

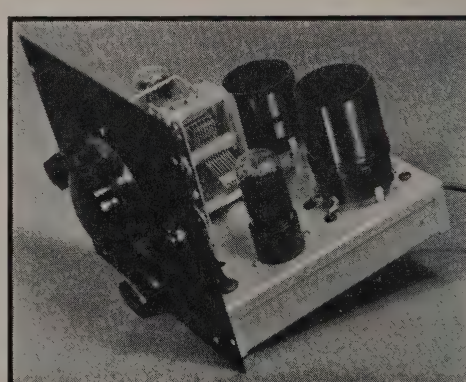
The low frequency coil has a plate winding of 85 turns, close-wound, and a tickler of approximately 65 turns, scramble-wound over a length of about $\frac{3}{8}$ inch at the ground end of the plate coil. The tickler is placed as close to the plate winding as is possible without danger of a short.

The tickler turns are rather critical, and for this reason only the approximate number is given; a different layout might call for 1 or 2 more or less turns. If too many tickler turns are provided, the oscillator will tend to motorboat or superregenerate over a part of the dial. If too few tickler turns are provided, the oscillator will not oscillate over the whole dial. The tickler polarity must be correct or oscillation will not be obtained.

When a check reveals that the tickler turns are correct, the tickler is treated with coil dope to hold the turns firmly in place.

The strength of the signal can be varied by opening or closing the cabinet lid, by changing

Figure 36.
TEST SIGNAL GENERATOR.
This signal generator covers the range from 440 to 1500 kc. in two jumps. A voltage regulator tube prevents change in plate voltage with line voltage fluctuations.



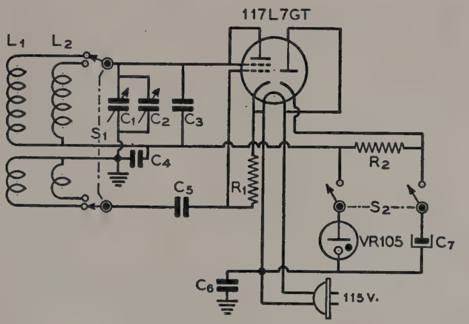


Figure 37.
WIRING DIAGRAM OF SIGNAL GENERATOR.

C_1 — 350- or 370-µfd. broadcast type condenser
C_2—2-µµfd. vernier (midget condenser with all but two double-spaced plates removed)
C_3 — 100-µfd. zero temperature coefficient padder
C_4 — .05-µfd. 400-volt tubular
C_5—250-µµfd. mica
C_6 — .05-µfd. 400-volt tubular
C_7 — 40-µfd. 150-volt electrolytic (225-volt peak)
R_1 — 50,000 ohms, ½ watt
R_2—1500 ohms, 2 watts
S_1—D.p.d.t. switch
S_2—D.p.s.t. switch
L_1, L_2 — Refer to text

the relative distance between signal generator and receiver or receiving antenna, etc.

When making use of a harmonic of the signal generator, it may be difficult to obtain sufficient signal without running an "antenna" wire inside the oscillator cabinet. If this is done, the vernier condenser should be adjusted *after* the pick-up wire is in place, as the presence of the wire will change the oscillator frequency slightly (as will opening or closing the cabinet lid).

The harmonics are sufficiently strong to be readily usable down to 8000 kc. when a "pick-up" wire is inserted inside the cabinet. Most heterodyne frequency meters operate on a fundamental frequency lower than 8000 kc., and no standard communications receiver employs an intermediate frequency below 440 kc. Hence, the frequency range is adequate for all normal purposes.

A.C.-D.C. Vacuum-Tube Voltmeter

The instrument illustrated in Figures 39, 41, and 42 is designed to permit measurement of a.c. and d.c. voltages on the ranges of 1, 10, 100, and 500 volts. The input resistance when measuring d.c. voltages remains constant at 100,000 ohms per volt on all ranges. The a.c. input impedance is approximately one-half the d.c. resistance, or 50,000 ohms per volt.

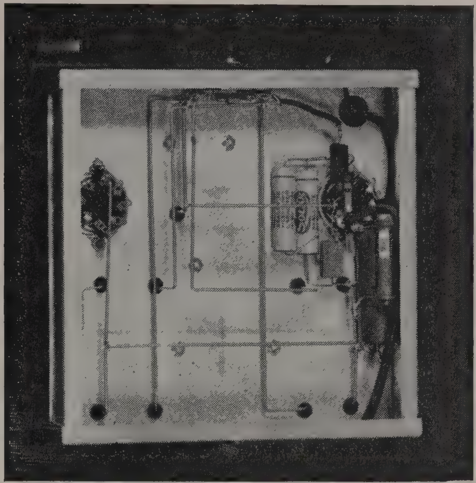


Figure 38.
UNDER-CHASSIS VIEW OF SIGNAL GENERATOR.

The circuit of the v.t.v.m. is shown in Figure 40. The triode section of the 6SQ7 acts as a linear d.c. voltmeter, while the diode section is used as a rectifier for a.c. voltages. For a.c. measurements, the rectified voltage is applied to the triode grid, while d.c. voltages to be measured are applied directly to the grid.

A bias cell in the grid lead is used as bias for the triode, and voltages to be measured are applied as additional negative bias. When switch S_1 is in its lowest position the voltage under measurement is applied directly to the grid, giving maximum sensitivity. With a 0–1 milliammeter at M, one volt d.c. at the grid will give full-scale indication; voltages down to .05 volt may be read quite easily on this scale. The additional multiplier resistors R_2, R_3, and R_4 allow full-scale indication at 10, 100, and 500 volts.

To measure a.c. voltages the "A.C." terminals are used. The a.c. voltage is rectified by the 6SQ7 diodes and applied across the grid-cathode circuit of the triode section as a d.c. voltage. Resistors R_2, R_3, and R_4 form the load circuit for the diode rectifier, while the input circuit is completed through the source of a.c. voltage under measurement.

For a.c. measurements, the v.t.v.m. has the same ranges as for d.c., but the *peak* values of the half-cycles are indicated by the meter. In most cases sine-wave a.c. will be measured, so the r.m.s. a.c. voltage will be given by multiplying the meter readings by .707. The full-scale r.m.s. readings for a.c. are .7, 7.07, 70.7, and 353.5 volts on the four ranges.

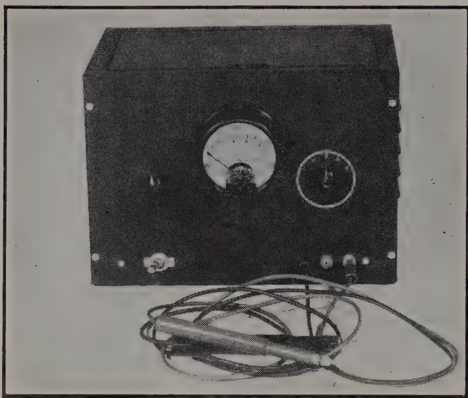


Figure 39.

A.C.-D.C. V.T. VOLTMETER.

The ranges for both a.c. and d.c. are 1, 10, 100, and 500 volts. The left knob operates the bridge-balancing control, and the one on the right operates the range switch.

Condenser C_2 is necessary to eliminate the effects of electrostatic pick-up from the power supply and nearby a.c. lines on the 10-volt scale.

The power supply is a simple affair employing a small b.c. type replacement transformer and a resistance-capacity filter. The filtering action is aided by the two voltage regulator tubes. These tubes hold the plate voltage constant at 255 volts, thus improving the accuracy under changes in line voltage.

As may be seen from the photographs, the power supply and voltage-regulator section is located at the rear of the chassis, while the 6SQ7 and its associated circuits occupy the section nearest the panel. The three input terminals are located at the lower right edge of the panel, with the power switch in a corresponding location at the left. The knob to the right of the meter operates the range-change switch, while the other knob is for balancing the meter bridge circuit. The cabinet measures 10 x 7 x 6 inches, and the chassis is 9 inches long and 5½ inches deep.

Calibration. Before attempting calibration, the v.t.v.m. power supply should be turned on and the meter adjusted to read zero by means of R_9. As the 6SQ7 warms up, the meter reading will change somewhat, but a stable condition should be reached in 2 or 3 minutes. After the unit has been warmed up, the range switch should be set to the 1-volt scale and a d.c. voltage of 1 volt applied to the d.c. terminals. The correct voltage may be obtained from a 1.4-volt flashlight cell by means of a potentiometer, an ordinary d.c. voltmeter being used to determine the correct setting of the potentiometer. If all goes well the meter will read exactly full scale with the 1 volt applied. If the meter should read too high or too low, the value of R_6 should be changed slightly, R_9 reset for balance with no input voltage, and the 1-volt indication again checked. The value of R_6 is not extremely critical, and it is quite probable that the value specified under the diagram will give satisfactory results.

After the instrument has been calibrated on the 1-volt scale, no further calibration is needed, if the multiplier resistors for the other

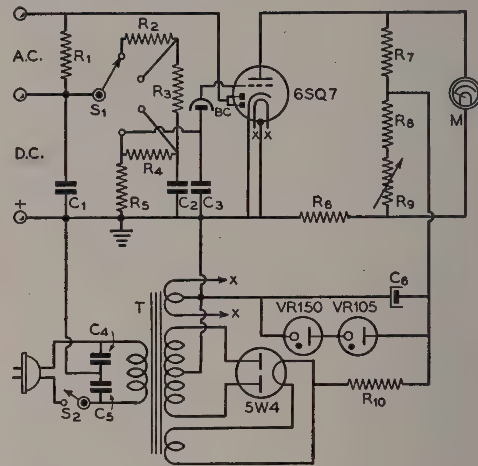


Figure 40.

WIRING DIAGRAM OF THE A.C.-D.C. VACUUM-TUBE VOLTMETER.

C_1—0.05-μfd. 600-volt tubular
C_2 — 0.1-μfd. 600-volt tubular
C_3 — 0.5-μfd. 600-volt tubular
C_4, C_5 — 0.05-μfd. 600-volt tubular
C_6—8-μfd. 450-volt electrolytic
R_1—1.0 megohm, 1 watt
R_2 — 40 megohms (4 10-megohm ½-watt in series)
R_3 — 9.0 megohms (5 megohms and 4 megohms ½-watt in series)
R_4—900,000 ohms (400,000 and 500,000 ohms ½-watt in series)
R_5—100,000 ohms, ½ watt
R_6 — 30,000 ohms, 1½ watts
R_7 — 10,000 ohms, 1½ watts
R_8 — 1500 ohms, 1½ watts
R_9—1000-ohm potentiometer
R_{10} — 2000 ohms, 10 watts
T — 580 c. t., 50 ma.; 5 v., 3 a.; 6.3 v., 2 a.
M—0-1 d. c. milliammeter
BC—1¼-volt bias cell
S_1—Single-pole 4-position switch
S_2 — S.p.s.t. toggle line switch

ranges are accurate. There is one type of measurement for which the v.t.v.m. *cannot* be used—it will be found impossible to get a correct reading on the 120-volt a.c. supply line, since condensers C_4 and C_5 place the chassis effectively at the center of the a.c. supply voltage, causing an incorrect reading.

High-Voltage Peak Voltmeter

A diode vacuum-tube voltmeter suitable for the measurement of high values of a.c. voltage is diagrammed in Figure 43. With the constants shown, the voltmeter has two ranges—500 and 1500 volts.

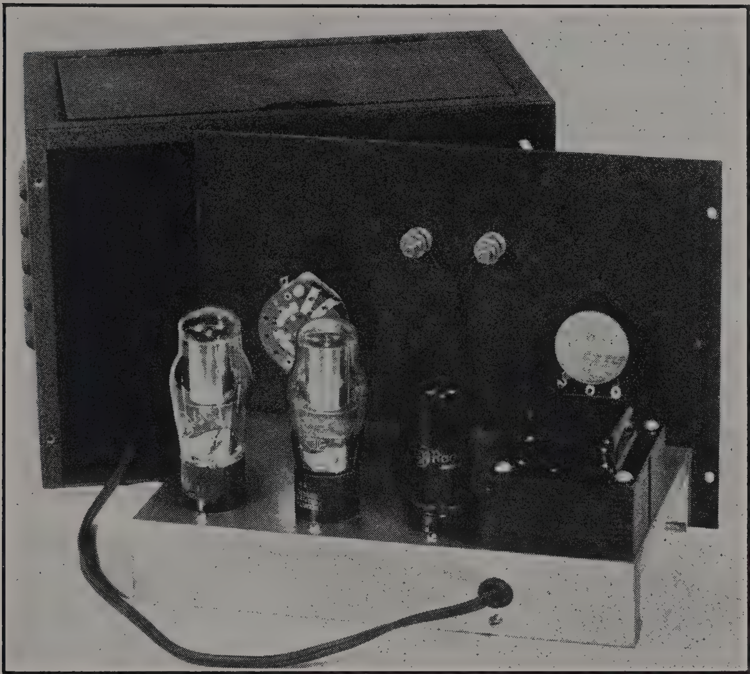


Figure 41.

V.T.V.M. REMOVED FROM ITS CABINET.

The power transformer, metal rectifier tube, and the two voltage regulator tubes are in a line, from right to left across the rear of the chassis. Rubber grommets insulate the meter terminals where they pass through the panel.

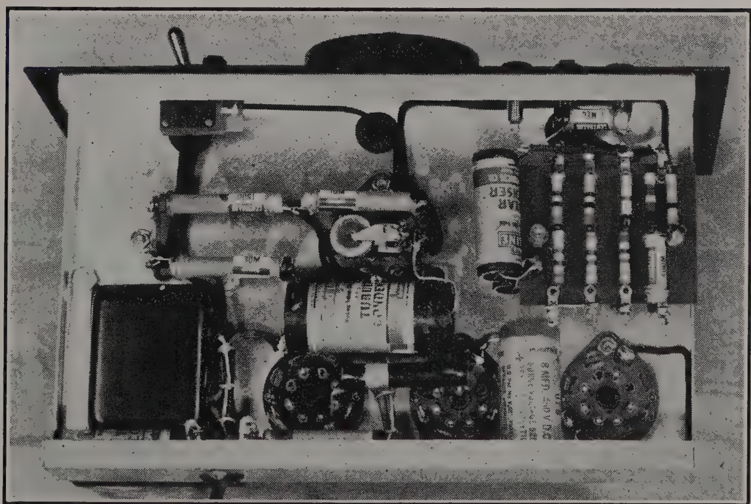


Figure 42.

UNDER-CHASSIS VIEW OF V.T.V.M.

The multiplier resistors are all located on the terminal plate seen at the upper right; a cabled set of leads goes through the chassis from the terminal strip to the range switch above deck. The bias cell holder is supported by having one of its terminals soldered directly to the grid connection on the 6SQ7 socket.

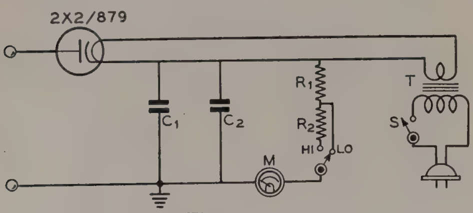


Figure 43.

WIRING DIAGRAM OF HIGH-VOLTAGE PEAK VOLTMETER.

C_1—.001-μfd. high-voltage mica	T—2.5 v., 1.75 a. filament transformer
C_2—1.0-μfd. high-voltage paper	M—0-1 d.c. milliammeter
R_1—500,000 ohms (2 0.25-megohm ½-watt in series)	S_{HI-LO} — S.p.d.t. toggle switch
R_2 — 1.0 megohm (4 0.25-megohm ½-watt in series)	S—S.p.s.t. toggle switch

Condensers C_1 and C_2 should be able to withstand in excess of the highest peak voltage to be measured. Likewise, R_1 and R_2 should be able to withstand the same amount of voltage. The easiest and least expensive way of obtaining such resistors is to use several low-voltage resistors in series, as shown in Figure 43. Other voltage ranges can be obtained by changing the value of these resistors, but for voltages less than several hundred volts a more linear calibration can be obtained by using a receiving-type diode. A calibration curve should be run to eliminate the appreciable error due to the high internal resistance of the diode, preventing the condenser from charging to the full peak value of the voltage being measured.

Wide Range Audio Oscillator

A source of variable-frequency audio frequency power having negligible harmonic content is of great usefulness in testing audio amplifiers and modulators. Such an oscillator is shown in Figures 44 to 46. It covers from 16.6 to 85,000 cycles, over which range the output remains substantially constant except at the high-frequency end of the range.

The satisfactory operation of the oscillator is dependent upon an automatic adjustment of the ratio between the regenerative and degenerative feedback between the output and the input circuit. The magnitude of the regenerative feedback is fixed and is fed back directly to the grid of the 6SJ7 through the resistance-capacity input circuit. The degenerative feedback is coupled through R_9 into the two lamps RL_1-RL_2 in the cathode circuit of the 6SJ7. These lamps have a very positive resistance-temperature characteristic, hence the amount of feedback voltage increases more rapidly than does the current through the lamps. The varying voltage drop across the lamp resistor circuit is determined by the change in magnitude of the audio current fed back from the plate circuit of the first 6V6-GT tube.

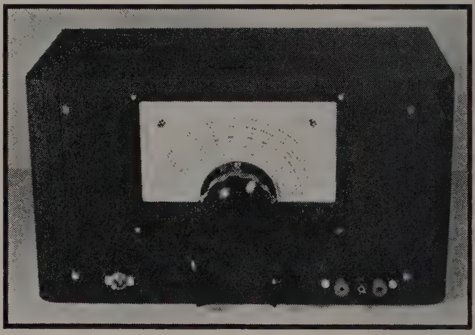
The output amplifier isolates the oscillator from the external circuit, allowing it to be fed into the primary of a transformer or into a low-impedance line without any reaction upon the oscillator caused by varying output circuit conditions. The voltage available from the low-impedance output terminals is variable up to about 1.25 volts over the audible range, dropping off slightly above 25,000 cycles. The high-impedance output terminals will supply about 18 to 20 volts.

The frame of the tuning condenser is at grid potential above ground. For this reason it has been found necessary to shield the entire oscillator portion of the unit from the power supply section, and, for that matter, from surrounding fields in general. The shielding was accomplished by placing a large shield directly behind the tuning condenser and between it and the power supply, and by inserting a small shield in an analogous position below the chassis. With these two shields in place, and with the entire unit in its shielding metal cabinet, there is no interaction between the oscillator and the line frequency, and no hum appears in the output. However, when the oscillator is removed from the cabinet the large area of the tuning condenser will pick up a certain amount of hum from surrounding a.c. lines. The result of this a.c. pickup will be

Figure 44.

WIDE RANGE AUDIO OSCILLATOR.

This unit covers the range from 16.6 to over 85,000 cycles; the output remaining substantially constant except at the high-frequency end of the range.



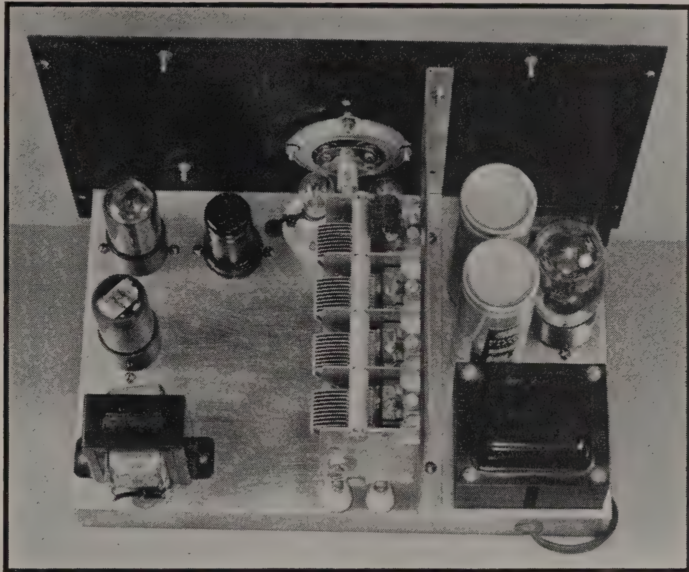


Figure 45.
PLACEMENT OF COMPONENTS ABOVE THE CHASSIS.
Note especially the metal shield, extending almost to the top of the panel, which separates the oscillator portion from the power supply.

quite noticeable on the lowest frequency scale, where the oscillator and its harmonics will tend to lock in with the various harmonics of the a.c. line frequency. For this reason, it is important that the oscillator be operated only when it is thoroughly enclosed in its shielding cabinet.

C_2 is used to maintain a capacity balance of the two sections, since the lower section has a much greater capacity to ground. It is placed across the upper half of the dual tuning condenser, the trimmers on the lower section are turned all the way out, and the trimmers in parallel with C_2 are varied until smooth oscillation is maintained over all bands. The adjustment is not particularly critical, although a cathode-ray oscilloscope will be of assistance in determining the best position. If the trimmers are not set correctly, it will be difficult to maintain oscillation in the vicinity of 150 cycles on the lowest scale, and there will be a peculiar dyssymmetry in the waveform in the vicinity of 1000 cycles on the second scale.

A range of about 10 to 1 will be covered on each set of resistors, and the coverages of these four ranges are as follows: 1.—16.6 to 150 cycles, 2.—150 to 1,150 cycles, 3.—1,500 to 10,000 cycles, 4.—10,000 to 80,000 cycles. These figures are the end calibration points of the various ranges. The frequency coverage actually is continuous from 16.6 cycles to over 85,000 cycles.

Calibration. The most satisfactory and least difficult method of calibration would be to check the unit by the zero-beat method against another audio oscillator which is already accurately calibrated. The unit also can be calibrated by means of an oscilloscope having a linear sweep oscillator going from about 10 to 5000 cycles, by utilizing the power line frequency as a base from which to start. The oscilloscope shown on page 521 was used in this case. The procedure, though simple, must be followed exactly as given in order that no error be introduced, since any error introduced at the outset would be cumulative throughout the calibration.

First, it is best to have both the phones and the vertical plates of the oscilloscope (through the amplifier in the 'scope) connected across the output of the oscillator. Then, with alternating current directly from the line fed into the horizontal plates of the 'scope, adjust the frequency of the oscillator on the lowest range until a figure such as is shown in Figure 47A is obtained. This figure indicates that the vertical deflection frequency is exactly twice that of the horizontal deflection frequency. If the local line frequency is 60 cycles, the oscillator is on 120 cycles.

Then turn in the tuning condenser until a circle is obtained on the screen of the 'scope—the audio oscillator is now operating on the *same* frequency as the local line. Now turn the condenser in still further until a figure the same as described in the preceding paragraph *but lying on its side* is obtained—the oscillator is now on *half* the frequency of the local line. If the oscillator condenser is turned in still

further it will be possible to obtain a figure which will have 1 loop in the horizontal plane and 3 loops in the vertical plane—the oscillator is then on one-third of the line frequency. These calibration points can then be marked on the proper dial scale by making a dot with a sharp pencil point inserted through the hole in the pointer (with the celluloid cover removed). In-between fractional ratios between the line frequency and the oscillator can be obtained for additional calibration points if Lissajou's figures, as described in the literature and in later paragraphs, are formed on the screen by careful adjustment of the oscillator frequency. The oscillator should now be returned to the position which gives the figure described in the preceding paragraph, with the oscillator on twice line frequency.

Turn the synchronization control until there is no interlocking between the incoming signal and the sweep oscillator, turn on the sweep oscillator, and adjust its frequency until a single stationary sine wave appears on the screen. The sweep oscillator is now on exactly 120 cycles. The standard of frequency has been transferred from the 60-cycle line to the 120-cycle linear sweep oscillator.

The determination of the calibration points for frequencies intermediate (fractional multiples) between the fundamental and integral harmonics, such as the second, third, etc., can be made through a knowledge of certain geometrical patterns which will be seen on the screen of the oscilloscope, called Lissajou's figures. Their interpretation is simple enough since they represent, when standing still, fractional relations between the two frequencies which are being impressed upon the vertical and horizontal plates of the oscilloscope.

The fractional relation between the two frequencies can be determined by a simple inspection of the waveform which appears on the 'scope. First, the number of complete bumps which appear along the top in the horizontal direction is counted; this is the numerator of the fraction. Then the number of traces is counted (this may be determined by counting the number of free "tails" at either end of the figure, or by taking *one more* than the number of crossovers on any ascension or descension of one-half of one of the sine waves) and this value is the denominator of the fraction which represents the relation between the frequency on the vertical plates with respect to the frequency of horizontal saw-tooth sweep.

Figure 47B shows a Lissajou's figure which represents a 4/3 ratio between the impressed voltage and the horizontal saw-tooth frequency. If the sweep frequency were still 120 cycles in this case, the input frequency would be 160 cycles. Calibration points for frequencies which are intermediate between integral multiples may be obtained in this way.

Now switch to the next higher frequency range and, keeping the *sweep* oscillator on 120 cycles, set the dial of the audio oscillator to the point where 2 complete sine waves appear on the screen. The oscillator will now be on 2/1 times 120 cycles, or 240 cycles. Put this down in the chart and increase frequency until 3 sine waves appear: this will be 3/1 or 360 cycles. Next come 4 sine waves or 480 cycles (the figure for this is shown in 47C), 5 sine

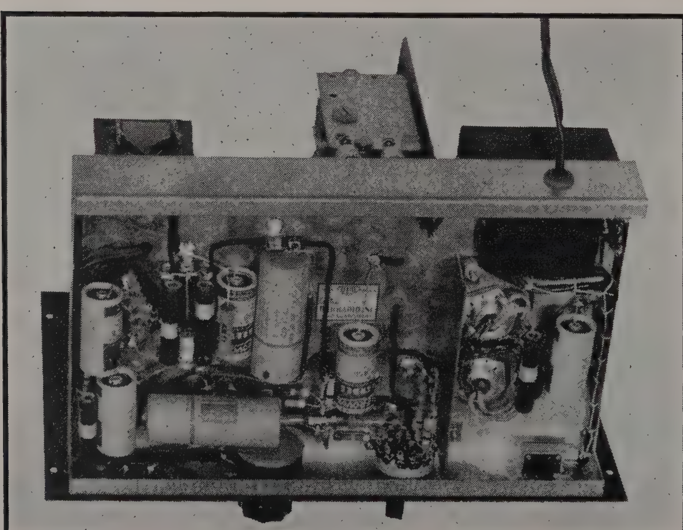


Figure 46.
UNDER-CHASSIS VIEW.
The small metal shield, between the range switch and resistors and the filter condenser leads in the power supply, is important if ripple difficulties are to be eliminated.

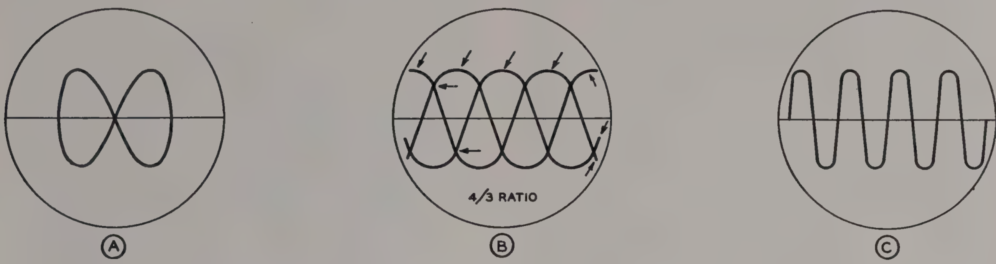


Figure 47. Examples of the oscilloscope patterns used in the calibration of the audio oscillator with the aid of a cathode-ray oscilloscope. (A) Pattern obtained with 60 cycles (sine wave a.c.) on the horizontal plates and 120 cycles (sine wave a.c.) on the vertical plates. (B) Figure showing a relation of 4/3 between the frequency of vertical deflection and the frequency of horizontal saw-tooth sweep. (C) Figure showing a relation of 4/1 between the vertical deflection and horizontal saw-tooth frequencies.

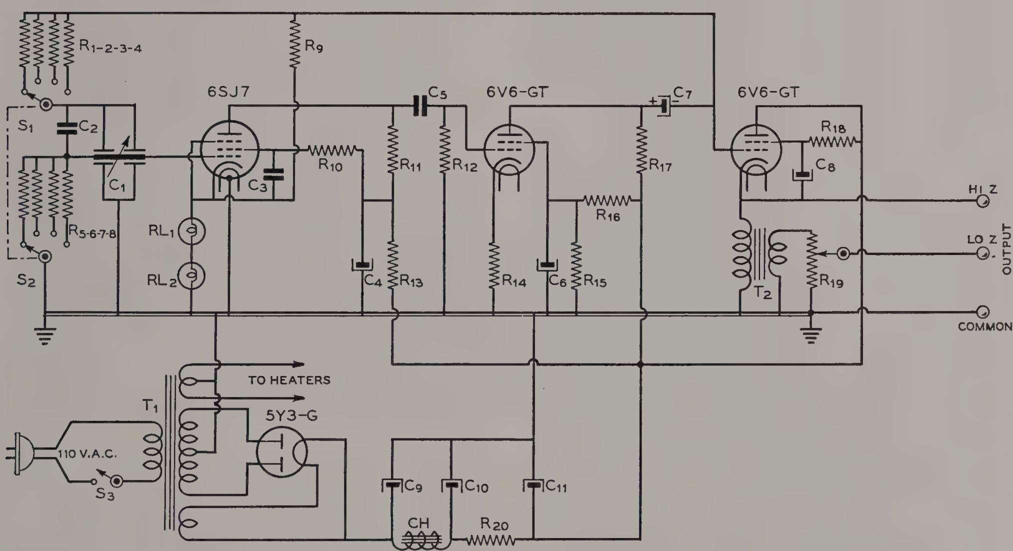


Figure 48.

WIRING DIAGRAM OF THE WIDE RANGE AUDIO OSCILLATOR.

C_1—4-gang 365-$\mu\mu$fd. b.c. condenser
C_2—.000075-μfd. midget mica
C_3—1.0-μfd. 400-volt tubular
C_4—8-μfd. 450-volt electrolytic tubular
C_5—1.0-μfd. 400-volt tubular
C_6, C_7, C_8, C_9—8-μfd. 450-volt electrolytic tubular
C_{10}, C_{11}—16-μfd. 450-volt electrolytic
R_1, R_5—10 megohms, ½ watt
R_2, R_6—1.25 megohms, ½ watt (1 megohm and 250,000 ohms in series)
R_3, R_7—150,000 ohms, ½ watt
R_4, R_8—20,000 ohms, ½ watt
R_9—2500 ohms, 1 watt
R_{10}—500,000 ohms, ½ watt
R_{11}—100,000 ohms, ½ watt
R_{12}—500,000 ohms, ½ watt
R_{13}—100,000 ohms, ½ watt
R_{14}—200 ohms, 10 watts
R_{15}—40,000 ohms, 1 watt
R_{16}—50,000 ohms, 1 watt
R_{17}—10,000 ohms, 10 watts
R_{18}—25,000 ohms, 10 watts
R_{19}—1000-ohm potentiometer
R_{20}—500 ohms, 10 watts
RL_1, RL_2—6-watt 120-volt tungsten mazda lamp
T_1—580 c.t., 50 ma.; 5 v. 3 a.; 6.3 v. 2 a.
T_2—Universal output to voice coil trans.
CH—10-hy. 65-ma. filter choke
S_1, S_2—2-pole 6-position switch (only 4 positions used)
S_3—S.p.s.t. a.c. line switch

waves or 600 cycles, 6 sine waves or 720 cycles, and 7 sine waves or 840 cycles.

Since it becomes difficult to count the number of waves accurately with a small c.r. tube, the standard frequency must be increased to enable the calibration of the higher ranges. This is a very interesting and comparatively simple procedure, but it must be followed carefully, step by step. First, the oscillator is tuned down in frequency again until there are 5 sine waves on the screen indicating 600 cycles. It is important in making all these adjustments to tune the oscillator carefully until the pattern stands quite still. Now retune the *sweep* oscillator in the oscilloscope until there are *6* sine waves on the screen where there were 5 before. The sweep is now on 100 cycles instead of 120—hence the 6 waves instead of the 5. Now retune the audio oscillator until there are *5* waves on the screen instead of 6, the oscillator now being on 500 cycles, and then retune the *sweep* oscillator until there is only 1 sine wave on the screen. This puts the sweep oscillator on 500 cycles, the new base frequency.

Now by switching the oscillator to the third range the frequencies of 1500, 2000, 2500, 3000, etc., may be checked by their multiple sine wave patterns. Then the audio oscillator may be shifted to 1000 cycles, the sweep oscillator shifted to 1000 cycles to give a single wave, and the frequencies from 3000 to 12,000 cycles checked.

For extremely high audio frequencies, the oscillator may be shifted to 5000 cycles and the sweep oscillator increased in frequency until a single sine wave is visible, showing that the sweep is on the same frequency. Then this frequency may be multiplied on up in the manner used for the lower frequencies until calibration up to 75,000 cycles or above is obtained.

As a check upon the entire calibration the entire process may be reversed and the difference between the resulting check line frequency and the actual frequency determined. If it is very far off, the whole process had better be repeated in order to obtain a more accurate calibration.

Cathode-Ray Oscilloscopes

Measurements of r.f. and a.f. voltage and wave form can easily be made with the aid of a cathode-ray oscilloscope. Such a device includes a vacuum tube which has two sets of deflecting plates for controlling a beam of electrons; this beam strikes a fluorescent screen on the face of the tube and traces a pattern of the signal applied to the control grid or deflection plates. The fluorescent screen in the tube produces a visual indication of the pattern of r.f. or audio voltages.

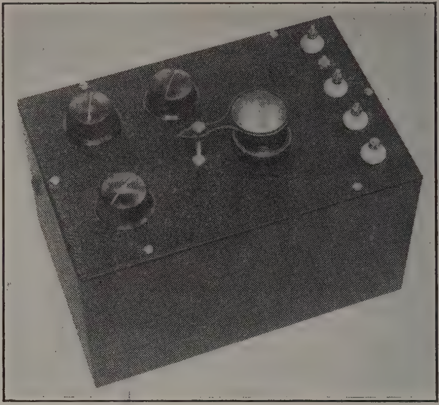


Figure 49.
CATHODE RAY MODULATION CHECKER.
This inexpensive oscilloscope is useful for obtaining trapezoidal modulation patterns. A cheap magnifying glass and tubular shade to keep out external light give a pattern comparable to that obtained with a 2-inch c.r. tube.

Some of the many uses of the cathode-ray oscilloscope in its various forms are as follows:

- Measurement of d.c. voltage or current.
- Measurement of peak a.c. and r.f. voltage.
- Trouble-shooting in receivers.
- Adjustment of i.f. stages (including band-pass).
- Measurement of audio amplifier distortion, overload and gain.
- Adjustment of phase-inversion circuits.
- Checking of power supplies.
- Checking of harmonic content.
- Measurement of phase angle and phase distortion.
- Measurements for dynamic tube characteristic curves.
- Checking of 'phone signals and per cent modulation by:
  - Modulation envelope
  - Trapezoidal pattern
  - Cat's eye pattern
- Making condenser power factor tests.
- Making overall frequency response tests.
- Determining unknown frequencies.
- Adjusting auto vibrators.
- Studying surges and transients.

Cathode-ray oscilloscopes are extremely useful for measuring percentage modulation and analyzing distortion in a 'phone transmitter.

While constructional data is given for two oscilloscopes, one a simple instrument for checking modulation in a radiophone transmitter and the other a more elaborate instrument possessing greater versatility, anyone contemplating construction of an oscilloscope should invest in one of the many excellent books on the subject, available very reasonably from *Rider, RCA Manufacturing Co., Dumont* and others. Because of space limitation, a comprehensive treatise on the theory, construction, and use of oscilloscopes is not within the scope of this book. This will be appreciated when it is realized that there appear books on oscilloscopes which contain over 100 pages devoted to applications of the instrument alone.

The accelerating anode potentials used in many oscilloscopes, particularly the larger sizes, are high enough to be very dangerous.

C. R. Modulation Checker

A very simple oscilloscope, such as the one shown in Figures 49 and 50, is entirely satisfactory for modulation checking. It consists of an RCA-913 cathode-ray tube which has a fluorescent screen approximately 1 inch in diameter. This tube, and a suitable power supply, are built into a small metal cabinet measuring 5 x 6 x 9 inches.

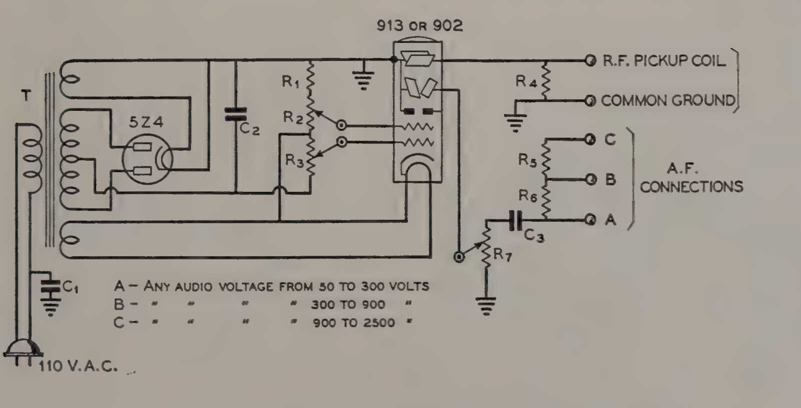
A dime magnifying glass obtainable at any five-and-ten-cent store gives a trapezoidal figure comparable in size to that of a 2-inch cathode-ray tube. The magnifying glass is held about 2 inches from the screen of the 913 by a piece of bakelite tubing which is slipped over the 913 and allowed to project slightly beyond the magnifying glass in order to keep out external light. If desired, a 902 (2-inch screen) may be substituted for the 913; no circuit changes will be required.

Three a.f. binding posts allow connection of the 'scope to the modulator of any 'phone transmitter with 5- to 1000-watts carrier power. No external coupling condenser is required; a lead may be connected directly to the class C amplifier plate return circuit at the modulation transformer terminals. *Beware of the high voltage.* Connections for a grid-modulated transmitter are similar, except that the modulation transformer connection is in the grid-return instead of the plate return circuit of the r.f. amplifier. The resistor network adapts the instrument for use on any transmitter at a moment's notice; no trouble will be experienced in getting just the right amount of audio deflecting voltage.

The network resistors R_5 and R_6 are not standard items; each is made up of 1-megohm 1-watt carbon resistors in series, R_5 requiring six such resistors and R_6 two. The 1-watt resistors are mounted on terminal strips.

When a voltage is applied to only one set of plates, a thin straight line is obtained on the face of the cathode-ray tube when the 25,000- and 50,000-ohm potentiometers are correctly adjusted.

When a modulated carrier voltage is applied to one set of plates, and the audio modulating voltage applied to the other, a *trapezoidal figure* will be produced during modulation. With 100 per cent modulation this pattern should be a straight-sided triangle, sharply



C_1—.01-μfd. tubular
C_2—1-μfd. 1000 volts w.v.
C_3—.01-μfd. mica, 5000 volts
R_1—150,000 ohms, 1 watt
R^2—50,000-ohm potentiometer
R_3—25,000-ohm potentiometer
R_4—5 meg., 1 watt
R_5—6 meg., 6 watts (see text)
R_6—2 meg., 2 watts (see text)
R_7—1-meg. pot.
T—350 v. each side c.t., 40 ma.

Figure 50.
WIRING DIAGRAM OF THE CATHODE RAY MODULATION CHECKER.

pointed. Typical patterns are shown for plate and grid modulation in the accompanying sketches, Figure 51.

The audio- or radio-frequency voltage should have an amplitude of at least 50 volts in order to cause good deflection on the screen. The amplitude should be sufficient to give a large pattern on the face of the tube. The 25,000- and 50,000-ohm potentiometers are adjusted to give sharp definition and a reasonable amount of illumination on the screen. The r.f. voltage can be secured by coupling a few turns of wire to the center of the modulated amplifier tank coil or to the antenna coupler.

The tube should not be allowed to run for more than an instant with no deflecting voltages applied, as a burned spot will appear on the screen of the tube if the electron stream is allowed to converge for very long on a single small spot on the screen.

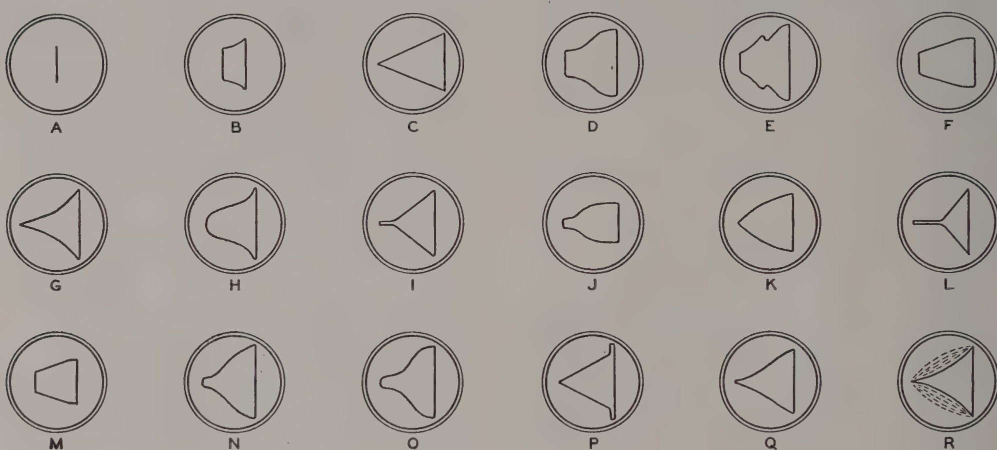


Figure 51.

TYPICAL OSCILLOGRAPHIC MODULATION PATTERNS.

It is assumed that there is negligible distortion in the a.f. voltage fed to the horizontal deflecting plates. (This voltage usually is taken from the last stage in the speech amplifier.) Also, except in the case of Figure R, it is assumed that there is negligible phase shift between the a.f. voltage applied to the horizontal deflecting plates and the a.f. voltage modulating the r.f. amplifier. Often an imperfect trapezoid at 100% modulation is a result of several factors, making it difficult to interpret the pattern and diagnose the particular trouble.

A: Unmodulated carrier signal.
B: Undistorted plate, grid, or cathode modulation, less than 100%.
C: Undistorted 100% plate modulation.
D: Plate modulation with inadequate or mismatched modulator.
E: Same as D with regeneration in modulated stage.
F: Plate modulated, insufficient grid excitation and/or bias to allow over 50% undistorted modulation. Grid modulated, too much excitation to allow over 50% modulation in upward direction.
G: Plate modulated, imperfect neutralization permitting regeneration.
H: Grid modulated phone with improper neutralization and reactive load.
I: Overmodulation of well designed, plate modulated transmitter. Too much audio input.
J: Grid modulation, excessive excitation or poor regulation of r.f. driver.
K: Insufficient excitation and bias on plate modulated zero bias (very high μ) triode.
L: Very bad overmodulation of plate modulated transmitter, resulting in serious clipping of negative peaks and bad splatter.
M: Maximum plate modulation of screen grid tube without screen modulation (screen bypassed for a.f.).
N: Suppressor modulated phone using separate r.f. driver, modulated approximately 100%.
O: Suppressor modulated 802 or 804 with crystal in grid circuit.
P: Parasitics in modulated amplifier, not present except on positive modulation peaks.
Q: Grid or cathode modulation, properly adjusted, approximately 100% modulation. Very little distortion.
R: Phase shift in speech system between point at which voltage is taken for horizontal deflection and the modulator output. No distortion.

Figure 52.
DE LUXE CATHODE RAY OSCILLOSCOPE.
With amplifiers and linear sweep, this oscilloscope can be built at a reasonable price yet will do almost anything a larger model will do. The cathode ray tube is a 2-inch RCA-902.

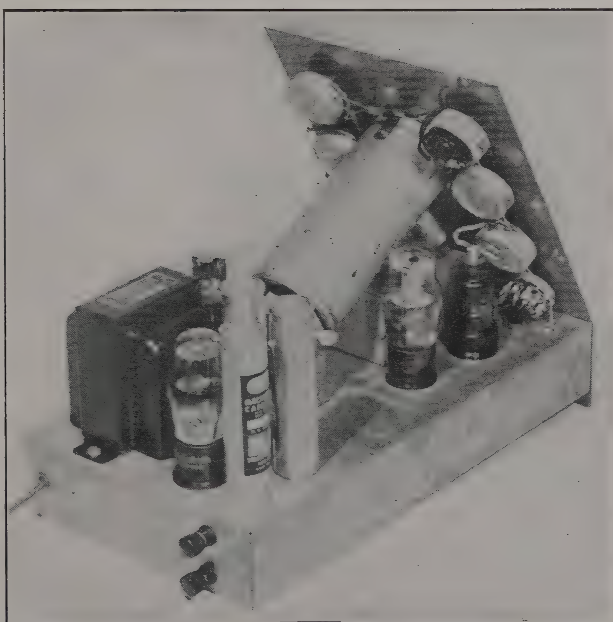
C.R. 'Scope with Sweep Circuit

Most audio-frequency measurements require a variable-frequency sweep oscillator circuit which can be synchronized with the frequency of the audio voltage being tested. For this purpose, a saw-tooth wave form is desirable; it can be obtained from a condenser-charge-and-discharge-circuit. The condenser is slowly charged, then rapidly discharged by means of a gas-filled type 885 triode which ionizes at a certain peak voltage and short-circuits the condenser in the plate circuit.

The sweep circuit oscillation can be synchronized with that of the audio-frequency signal by applying a small portion of the latter to the grid circuit of the type 885 tube. The approximate frequency of the saw-tooth oscillator is adjusted by means of the capacity in the plate circuit of the tube and the value of the resistance in series with the B-plus lead from this tube. The output of this oscillator must be amplified by a high-gain audio stage in order to provide sufficient voltage to produce a sweep across the screen of the cathode-ray tube.

The oscilloscope diagrammed in Figure 54 contains vertical and horizontal deflection plate amplifiers, linear sweep and most of the adjuncts found in the most expensive oscilloscopes. The only difference is the use of a small 902 2-inch cathode ray tube for the sake of economy.

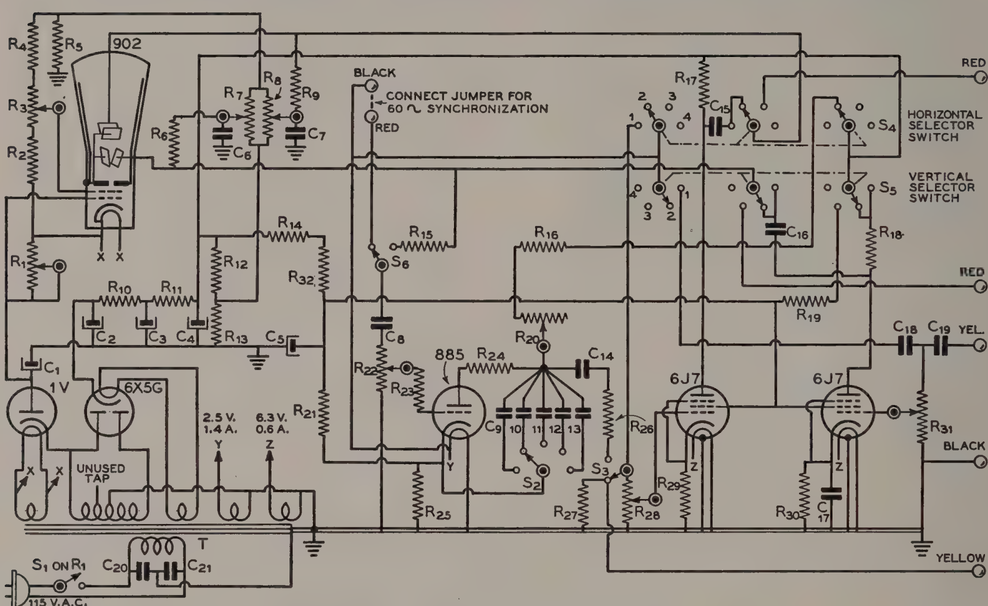
Figure 53.
INTERIOR CONSTRUCTION OF THE C.R. OSCILLOSCOPE.
The piece of heavy iron pipe shields the cathode ray tube both inductively and electrostatically. The pipe is supported from the panel by angle brackets. Sufficient space should be allowed to rotate the power transformer to eliminate any inductive coupling remaining after the pipe shield has been installed.



Construction. The galvanized iron pipe seen in Figure 53 acts as a shield over the c.r. tube, greatly reducing both inductive and electrostatic pickup of a.c. ripple. The pipe is just sufficiently large to take the tube. Further reduction in ripple is obtained by rotating the power transformer slightly until the sharpest line on the tube screen is obtained.

It is important that the various leads to the potentiometers and switches *not* be cabled to-

Figure 54.
VERSATILE CATHODE RAY OSCILLOSCOPE INCORPORATING DEFLECTION PLATE AMPLIFIERS, LINEAR SWEEP, AND 902 2-INCH C.R. TUBE.



R_1—25,000-ohm potentiometer with a.c. line switch
R_2—25,000 ohms, ½ watt
R_3—50,000-ohm potentiometer
R_4, R_5 — 100,000 ohms, 1 watt
R_6—2 megohms, ½ watt
R_7, R_8 — 100,000-ohm potentiometers
R_9—4 megohms, ½ watt
R_{10}, R_{11} — 2500 ohms, 1 watt
R_{12}—250,000 ohms, 1 watt
R_{13}—50,000 ohms, ½ watt
R_{14}—40,000 ohms, 1 watt
R_{15} — 2 megohms, ½ watt
R_{16}—750,000 ohms, ½ watt
R_{17}, R_{18} — 100,000 ohms, 1 watt
R_{19}—200,000 ohms, 1 watt
R_{20}—5-megohm potentiometer
R_{21}—40,000 ohms, 1 watt
R_{22} — 50,000-ohm potentiometer
R_{23}—25,000 ohms, ½ watt
R_{24}—200 ohms, ½ watt
R_{25}—1500 ohms, ½ watt
R_{26}, R_{27} — 1 megohm, ½ watt
R_{28}—3-megohm potentiometer
R_{29}, R_{30} — 1000 ohms, ½ watt
R_{31}—500,000-ohm potentiometer
R_{32}—40,000 ohms, 1 watt
C_1, C_2, C_3, C_4—8-μfd. 450-volt electrolytics
C_5—2-μfd. 200-volt electrolytic
C_6, C_7, C_8—0.1-μfd. 400-volt tubular
C_9 — 0.5-μfd. 400-volt tubular
C_{10}—0.1-μfd. 400-volt tubular
C_{11}—.02-μfd. 400-volt tubular
C_{12}—.005-μfd. mica
C_{13}—.001-μfd. mica
C_{14}, C_{15}, C_{16}—0.1-μfd. 400-volt tubular
C_{17}—.05-μfd. 400-volt tubular
C_{18}, C_{19}, C_{20}, C_{21}—0.1-μfd. 400-volt tubular
S_1—Line switch on R_1
S_2—5-position single-pole switch
S_3—S.p.d.t. toggle switch
S_4, S_5—3-circuit 4-position, non-shorting switch
S_6—S.p.d.t. toggle switch
T—Cathode ray oscilloscope power transformer
C.r. tube—902
C.r. tube mounting—Amphenol 913 plug and bracket assembly.

gether for the sake of appearance. The input circuits to the deflection plate amplifiers and the synchronizing voltage input circuit of the 885 saw-tooth oscillator should be isolated as much as possible from each other and from other parts of the instrument. The use of shielded leads will facilitate this.

The power supply and c.r. tube circuits should be wired first, and this wiring should be checked before proceeding with the wiring of the amplifiers and saw-tooth oscillator. In making this check the intensity control is left at maximum resistance until the tubes are hot, otherwise the screen may be either temporarily or permanently damaged. The intensity and focussing controls (R_1 and R_3, respectively) are varied until a fine spot appears on the screen. This spot should *never* be made any brighter than necessary, nor allowed to remain stationary except for brief periods, as it will burn a dead spot on the screen.

If the tube fails to operate the rectifier circuit is apt to be at fault. The required anode no. 1 voltage is *negative* with respect to ground, and the rectifier tubes used have cathodes.

If these circuits are satisfactory, the balance of the instrument can be wired. The adherence to the specified values of the components is quite important, as these values are rather critical if proper performance is to be obtained. The horizontal amplifier cathode resistor is intentionally not by-passed, to avoid bad tails on the sweep, and to preserve the sweep linearity.

Looking at Figure 52, the controls on the left side of the panel are as follows, reading from top to bottom:

Intensity and a.c. switch (R_1 and S_1),
Fine sweep frequency (R_{20}),
Horizontal amplifier gain (R_{28}),
Horizontal selector switch.

On the right side, reading from top to bottom:

Focussing (R_{22}),
Synchronizing (R_{22}),
Vertical amplifier gain (R_{31}),
Vertical selector switch.

The course sweep frequency control is in the center at the bottom. The two toggle switches are S_3, which connects the input of the horizontal amplifier to either the saw-tooth sweep oscillator or to an external deflecting voltage, and S_6, which allows the saw-tooth sweep to be synchronized with either the line frequency or with the deflecting voltage being applied to the vertical plates. Ground connections can be made to either of the two bottom binding posts (marked "Black" on the diagram). The posts directly above are marked "Yellow," and are used when the deflecting voltages are to be applied through the amplifiers. For connecting directly to the deflecting plates in the c.r. tube, the inner two posts are used (marked "Red"). R_7 and R_8 are the vertical and horizontal spot positioning controls, respectively, and are located at the rear of the chassis.

Various circuit arrangements can be obtained by changing the position of the horizontal selector switch. These are listed below, the positions referring to the numbers on the diagram, Figure 54.

Pos. 1—External voltage applied to the deflecting plates through the amplifier, or line frequency sweep voltage applied to the plates through the amplifier.
Pos. 2—Saw-tooth sweep voltage applied to the deflecting plates through the amplifier.
Pos. 3—External voltage applied directly on the deflecting plates.
Pos. 4—Off.

Similar conditions are obtainable in the case of the vertical deflection circuits by selecting various positions of the vertical selector switch, as follows:

Pos. 1—Line frequency sweep voltage applied to the deflecting plates through the amplifier.
Pos. 2—External voltage applied to the plates through the amplifier.
Pos. 3—External voltage applied directly on the deflecting plates.
Pos. 4—Off.

Workshop Practice

With a few possible exceptions, such as fixed air condensers and wire-wound transmitting coils, it hardly pays one to attempt to build the components required for the construction of an amateur transmitter. This is especially true when the parts are of the type used in construction and replacement work on broadcast receivers, as mass production has made these parts very inexpensive.

Transmitters. Those who have and wish to spend the necessary time can effect considerable monetary saving in their transmitters by building them from the component parts. The necessary data are given in the construction chapter of this handbook.

To many builders, the construction is as fascinating as the operation of the finished transmitter; in fact, many amateurs get so much satisfaction out of building a well-performing piece of equipment that they spend more time constructing and rebuilding equipment than they do operating the equipment on the air.

Those who are not mechanically minded and are more interested in the pleasures of working dx and rag chewing than in experimentation and construction will find on the market many excellent transmitters which require only line voltage and an antenna. If you are one of those amateurs, you will find little to interest you in this chapter.

Receivers. There is room for argument as to whether one can save money by constructing his own communications receiver. The combined demand for these receivers by the government, amateurs, airways, short-wave listeners, and others has become so great that it may be argued that there is no more point in building such a receiver than in building a regular broadcast set. Yet, many amateurs still prefer to construct their own receivers—in spite of the fact that it costs almost as much to build a receiver as to purchase an equivalent factory-made job—either because they enjoy construction work and take pride in the fruits of their efforts, or because the receiver must meet certain specifications and yet cost as little as possible.

The only factory-produced receiver that is sure to meet the requirements of every amateur or short-wave listener is the rather expensive de luxe type having every possible refinement. An amateur of limited means who is interested only in c.w. operation on two or three bands, for instance, can build himself, at a fraction of the cost of a de luxe job, a receiver that will serve his particular purpose just as well. In the receiver construction chapter are illustrated several relatively inexpensive receivers which, for the particular purpose for which they were designed, will perform as well as the costliest factory-built receiver.

Types of Construction

Breadboard. The simplest method of constructing equipment is to lay it out in breadboard fashion, which consists of fastening the various components to a board of suitable size with wood screws or machine bolts, arranging the parts so that important leads will be as short as possible.

While this type of construction is also adaptable to receivers and measuring and monitoring equipment, it is used principally for transmitter construction, and remains a favorite of the c.w. amateurs using high power.

Breadboard construction requires a minimum of tools; apparatus can be constructed in this fashion with the aid of only a rule, screwdriver, ice pick, saw, and soldering iron. A hand drill will also be required if it is desired to run part of the wiring underneath the breadboard, or if bolts are used to fasten down the parts. Ordinary carpenter's tools will be quite satisfactory.

Danger from accidental electrical shock usually is greatest with this type construction because of the exposed components.

Metal Chassis. Though quite a few more tools and considerably more time will be required for its construction, much neater equip-

ment can be built by mounting the parts on sheet metal chassis instead of breadboards. This type of construction is advisable when shielding of the apparatus is necessary, as breadboard construction does not particularly lend itself to shielding. The appearance of the apparatus may be further enhanced by incorporating a front panel upon which the various controls are placed. A front panel minimizes the danger of shock.

If sufficient pains are taken with the construction, and a front panel is used in conjunction with either a dust cover (cabinet) or enclosed relay rack, the apparatus can be made to resemble or even to rival factory-built equipment in appearance.

Dish type construction is practically the same as metal chassis construction, the main difference lying in the manner in which the chassis is fastened to the panel. Examples of both types are shown in Figure 1.

Special Frameworks. For high-powered r.f. stages, many amateur constructors prefer to discard the more conventional types of construction and employ instead special metal frameworks and brackets which they design specially for the parts which they intend to use. These are usually arranged to give the shortest possible r.f. leads and to fasten directly behind a relay rack panel by means of a few bolts, with the control shafts projecting through corresponding holes in the panel.

Tools

Beautiful work can be done with metal chassis and panels with the help of only a few inexpensive tools. However, the time required for construction will be greatly reduced if a fairly complete assortment of metal-working tools is available. Thus, it can be seen that while an array of tools will speed up the work, excellent results may be accomplished with but few tools, if one has the time and patience.

The investment one is justified in making in tools is dependent upon several factors. If you like to tinker, there are many tools useful in radio construction that you would probably buy anyway, or perhaps already have, such as screwdrivers, hammer, saws, square, vise, files, etc. This means that the money taken for tools from your radio budget can be used to buy the more specialized tools, such as socket punches or hole saws, taps and dies, etc.

The amount of construction work one does determines whether buying a large assortment of tools is an economical move. It also determines if one should buy the less expensive type offered at surprisingly low prices by the familiar mail order houses, "five and ten" stores and chain auto-supply stores, or whether one

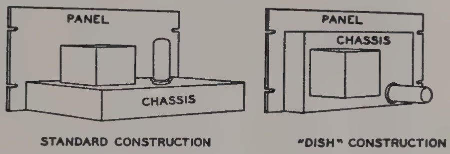


Figure 1.
TWO TYPES OF RACK-AND-PANEL CONSTRUCTION.

should spend more money and get first-grade tools. The latter cost considerably more and work but little better when new, but will outlast several sets of the cheaper tools. Therefore they are a wise investment for the experimenter who does lots of construction work (if he can afford the initial cash outlay). The amateur who constructs only an occasional piece of apparatus need not be so concerned with tool life, as even the cheaper grade tools will last him several years, if they are given proper care.

The hand tools and materials in the accompanying lists will be found very useful around the home workshop. Materials not listed but ordinarily used, such as paint, can best be purchased as required for each individual job.

ESSENTIAL HAND TOOLS AND MATERIALS

1 *Good* electric soldering iron, about 100 watts, with "radio" tip
1 Spool rosin-core wire solder
1 Jar soldering paste (non-corrosive)
1 Each large, medium, small, and midget screwdrivers
1 *Good* hand drill (eggbeater type), preferably two speed
1 Pair regular pliers, 6 inch
1 Pair long nose pliers, 6 inch
1 Pair cutting pliers (diagonals), 5 inch or 6 inch
1 $1\frac{1}{8}$ inch tube-socket punch
1 "Boy Scout" knife
1 Combination square and steel rule, 1 foot
1 Yardstick or steel pushrule
1 Scratch awl or ice pick scribe
1 Center punch
1 Dozen or more assorted round shank drills (as many as you can afford between no. 50 and $\frac{1}{4}$ or $\frac{3}{8}$ inch, depending upon size of hand drill chuck)
1 Combination oil stone
Light machine oil (in squirt can)
Friction tape

HIGHLY DESIRABLE HAND TOOLS AND MATERIALS

1 Good ball peen hammer, $\frac{3}{4}$ or 1 pound

1 Hacksaw with coarse and fine blades, 10 or 12 inch
1 Bench vise (jaws at least 3 inch)
1 Spool plain wire solder
1 Carpenter's brace, ratchet type
1 Square-shank countersink bit
1 Square-shank countersink
1 Square-shank taper reamer, small
1 Square-shank taper reamer, large (The two reamers should overlap; ½ inch and ⅞ inch size will usually be suitable.)
1 ⅞ inch tube-socket punch (for electrolytic condensers)
1 Square-shank adjustable circle cutter for holes to 3 inch
1 Set small, inexpensive, open-end wrenches
1 Pair tin shears, 10 or 12 inch
1 Cold chisel (½ inch tip)
1 Wood chisel (½ inch tip)
1 Pair wing dividers
1 Coarse mill file, flat, 12 inch
1 Coarse bastard file, round, ½ or ¾ inch diameter
6 or 8 Assorted small files: round, half-round, triangular, flat, square, rat-tail.
4 Small "C" clamps
Steel wool, coarse and fine
Sandpaper and emery cloth, coarse, medium, and fine
Rubber cement
File card or stiff brush

USEFUL BUT NOT ESSENTIAL TOOLS AND MATERIALS

1 Cheap carpenter's claw hammer
1 Jig or scroll saw (small) with metal-cutting blades
1 Small wood saw (crosscut teeth)
1 Each square-shank drills: ⅜, 7/16, and ½ inch
1 Tap and die outfit for 6-32 and 8-32 machine screw threads. (A complete set is not necessary, as other sizes seldom will be needed.)
4 Medium size "C" clamps
Lard oil (in squirt can)
Kerosene
Duco or polystyrene cement (coil dope)
Empire cloth
Alcohol
Clear lacquer ("industrial" grade)
Lacquer thinner
Dusting brush
Paint brushes
Sheet celluloid, Lucite, or polystyrene
Acetone
1 Carpenter's plane, 8 inch or larger
1 Metal punch
1 Each "Spintite" wrenches, ¼ and 5/16 inch to fit standard 6-32 and 8-32 nuts used in radio work and two common sizes of Parker Kalon metal screws

The foregoing assortment assumes that the constructor does not want to invest in the more expensive power tools, such as drill press, grinding head, etc. If power equipment is purchased, obviously some of the hand tools and accessories listed will be superfluous. A drill press greatly facilitates construction work, and it is unfortunate that a good one costs as much as a small transmitter. A booklet* available from the Delta Manufacturing Co. will be of considerable aid to those who have access to a drill press.

Not listed in the table are several special-purpose radio tools which are somewhat of a luxury but are nevertheless quite handy, such as various around-the-corner screwdrivers and wrenches, special soldering iron tips, etc. These can be found in the larger radio parts stores and are usually listed in their mail order catalogs. It is not uncommon to find amateurs who have had sufficient experience as machinists to design and produce tools for special purposes.

Tool Hints

Of equal importance in maintaining one's supply of necessary tools and assorted materials is the assignment of each tool to one particular location. The greatest loss of time in any shop is usually incurred by searching for tools which are not in their proper place.

Amateurs in or near the larger cities will often find it profitable to visit that section of the city where may be found many large stores that deal in used machinery and tools. It is quite commonplace to find used tools of high quality in good condition at a low price.

Vises. A vise is one of the few things that can be bought second-hand with safety. Brief inspection should settle its fitness: examine the screw, check for wobble and parallel jaws. A used machinist's vise is much better than a new, wobbly "home mechanic's vise." Three-inch jaws are usually large enough for radio work. If the vise is mounted on a corner of the bench a swivel is unnecessary, saving considerable expense. Only a few vises are sold with copper jaw plates, but these easily can be made as shown in Figure 2. It is most essential that a vise constantly used on bakelite, brass, and aluminum does not bite into the work with its steel jaws.

Soldering Irons. A prerequisite to a good soldering job is a good iron. If one can afford two irons, a 150-watt size for heavy work and a smaller 75-watt size for light work and getting into tight places are highly desirable. How-

* "Getting the Most Out of Your Drill Press," James Tate, Delta Manufacturing Company, Milwaukee, Wisconsin.

ever, a single 100-watt iron will do nicely for most purposes.

Do not get a high wattage iron that is relatively small physically. Such an iron must be used continuously to keep it from becoming too hot. When such an iron is left plugged in and is not used for several minutes, the iron will become so hot that it will curdle the solder adhering to the tip, making frequent filing and retinning necessary. An aluminum rest which presents considerable surface to the air and to the iron will prevent an iron from becoming overheated when not in actual use. Such a heat-dissipating rest for the iron can be made from an old aluminum automobile piston by sawing it off diametrically at the center of the wrist pin hole.

For occasional extremely heavy work, a soldering copper such as is used by sheet metal workers (heated in a gas flame) will be found very useful. This type of iron is merely a heavy copper tip fastened to a steel rod which has a wooden handle. Since the mass of the tip is great, it will hold heat for a long time, and is just the thing for working on large, heavy gauge subpanels, and on antennas where no current is available for an electric iron nearby. If heated sufficiently, they can be carried for considerable distance before becoming too cold for satisfactory soldering work.

An alternative for soldering joints at a distance from an a.c. power source is a small alcohol torch, obtainable for about 75 cents.

Wood Saws. There are many types of wood saws on the market, but for amateur construction work those listed in the tool tables are usually sufficient. Saws will work much better and last much longer if properly cared for. Keep the blades in good shape by smearing them with a thin film of vaseline after they are used. A rusty saw will not do good work. When it becomes necessary, as it does from time to time, to have them sharpened, let a good joiner do the work; it is a job for an expert, usually available in local hardware stores.

Metal Saws. The hacksaw has become an almost universally standardized tool for the amateur workshop. The replaceable blades are obtainable with varying numbers of teeth. The coarse blades, having 14 or 18 teeth per inch, can be used for bakelite or ebonite; for most metals a medium tooth blade with about 22 teeth per inch is desirable; and for very thin sheets a blade having 32 teeth per inch is best. Ordinarily, the harder the metal, the finer the blade that should be used.

When replacing saw blades, keep in mind that hacksaw blades should be put in place with the teeth pointing *towards the tip* of the saw, while jig or scroll saw blades should have the teeth *towards the handle,* in order to keep the work pressed on the cutting bench.

Files. When using a file, the handle should be grasped firmly in the right hand, with the thumb on top, and the left hand should rest on the tip to guide it. The pressure of the left hand on the file should be eased off and that of the right hand increased as the file proceeds across the work. The return stroke should be made with a minimum amount of pressure, or better, with the file raised from the work. The file should be cleaned often, both during and after work, in order to remove the filings which stick to the teeth. These may scratch the work if allowed to remain. A "file card" is inexpensive and will remove the burrs quickly.

To Resharpen Old Files. Wash the files in warm potash water to remove the grease and dirt, then wash in warm water and dry by heat. Put 1½ pints warm water in a wooden vessel, put in the files, add 3 ounces blue vitriol finely powdered, and 3 ounces borax. Mix well and

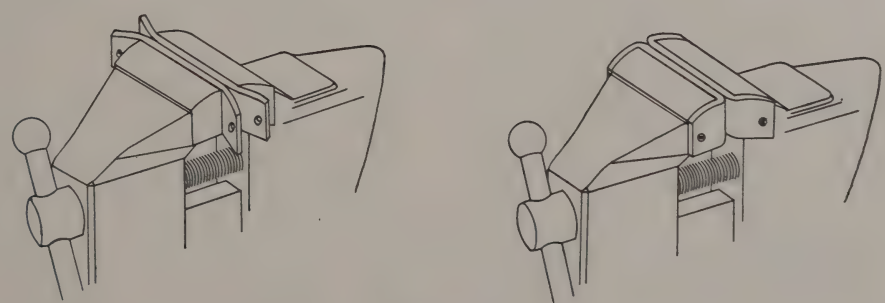


Figure 2.
INSTALLATION OF SOFT JAW PLATES.
The strips of copper are clamped in the vise, and the ends are then bent over and attached to the vise jaws.

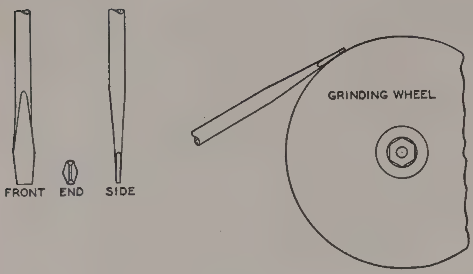


Figure 3.
SHARPENING A SCREWDRIVER.
The way to hold the screwdriver blade on the grinding wheel while sharpening it, and the appearance of a properly sharpened blade.

turn the files so that every one may come in contact with the mixture. Add 10½ ounces sulphuric acid and ½ ounce cider vinegar. Remove the files after a short time, dry, rub with olive oil, wrap in porous paper. Coarse files should be kept in the mixture for a longer time than fine ones. Be careful not to allow the solution to get on the hand.

File Lubricant. When filing aluminum, dural, etc., the file should be oiled or rubbed in chalk, but will cut slower than with no lubricant. However, the file will last much longer.

Screwdrivers. To do good work, several sizes of screwdrivers are necessary. There should be a blade to fit each of the screwheads in common use. The length of the slot in the screw-head and the width of the screwdriver blade should be as near alike as possible. If the blade is too narrow it will be twisted when the screw is tightened, and if it is too wide the screw is apt to be burred and the work scratched. Remember, also, that the thickness of the blade varies directly with the width. For best results, the blade should fit the slot snugly for its full width. The length of the blade is determined chiefly by the accessibility of the work, and to some extent by the choice of the worker. It should not be longer than necessary. A complete set of high quality screwdrivers with plastic handles (*Stanley*), which are excellent for radio and electrical work, can be obtained from hardware dealers for about $2.50.

Screw Lubricant. Put hard soap on lag screws, wood screws, or any screw for wood. It will surprise you how much easier they will turn in. The soap also will prevent, or at least reduce, splitting.

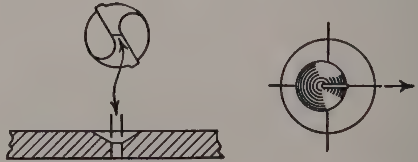
Power Drills and Drilling. Although most of us do not so consider it, a twist drill is nothing more than a modified jackknife. It has a cutting edge, an angle of clearance, and an angle of rake, just as has a jackknife (or a lathe tool). The technique to be followed in drilling is, therefore, a function of the type of material worked on, as well as the speed and accuracy desired. As the drill proceeds, a chip cut out is above the cutting edge. This has the effect of pulling the drill farther into the material. This is determined by the "angle of rake" and the hardness of the material. If the material is hard at the point of cutting, the resistance to downward motion here is great enough to overcome the pulling effect of the angle of rake.

For steel or iron, the shavings which come out of the hole around the drill should be spiral and continuous. For softer metals, especially brass, the drill should have no angle of rake (or lip, as the forward projecting cutting edge is called). The shavings for this drill will be small chips. If this shape drill is not used, the drill will feed into the metal very rapidly and will usually jam.

As the tip of a drill is not a point, but a straight line perpendicular to its major axis, a drill will usually waltz all over before it starts to drill unless a guide hole is punched at the point you wish to drill. The maximum diameter of this hole should be at least equal to the width of the drill tip. A center punch impression will suffice for small holes. With drills of over ¼-inch diameter, this method of starting the drill is usually impractical, as the diameter of the center-punch hole is prohibitively large. This difficulty is avoided by first drilling a smaller guide hole which can be started with a center-punch hole.

A great deal could be said about drilling speeds and feeds, but it would be of little value to the average person. Just remember that in drilling steel or iron, the drill point should be well lubricated with lard oil or a medium grade of machine oil. The weight of oil commonly used in oiling lawn mowers is about correct. This serves a double function for most machin-

Figure 4.
STARTING THE DRILL.
A drill should always be started with a center punch impression or small hole that is at least as large as the flat portion of the drill tip. If the guide hole is a little off, the drill can be "fudged over" by means of a chisel mark as shown to the right.



ists. The first, of course, is lubrication. The second is to keep the work cool. The oil flows from hot points to cold ones more quickly than heat flows from the hot points of the drill to the cooler ones. But, for amateur use, the oil assumes a third role: that of a temperature indicator. The oil should never evaporate visibly to form a cloud around the work (this vapor looks like steam).

NUMBERED DRILL SIZES

DRILL NUMBER	Diameter (in.)	Clears Screw	Correct for Tapping Steel or Brass †
1	.228	——	——
2	.221	12-24	——
3	.213	——	14-24
4	.209	12-20	——
5	.205	——	——
6	.204	——	——
7	.201	——	——
8	.199	——	——
9	.196	——	——
10*	.193	10-32	——
11	.191	10-24	——
12*	.189	——	——
13	.185	——	——
14	.182	——	——
15	.180	——	——
16	.177	——	12-24
17	.173	——	——
18*	.169	8-32	——
19	.166	——	12-20
20	.161	——	——
21*	.159	——	10-32
22	.157	——	——
23	.154	——	——
24	.152	——	——
25*	.149	——	10-24
26	.147	——	——
27	.144	——	——
28*	.140	6-32	——
29*	.136	——	8-32
30	.128	——	——
31	.120	——	——
32	.116	——	——
33*	.113	4-36 4-40	——
34	.111	——	——
35*	.110	——	6-32
36	.106	——	——
37	.104	——	——
38	.102	——	——
39*	.100	3-48	——
40	.098	——	——
41	.096	——	——
42*	.093	——	4-36 4-40
43	.089	2-56	——
44	.086	——	——
45*	.082	——	3-48

† Use next size larger drill for tapping bakelite and similar composition materials (plastics, etc.).

*** Sizes most commonly used in radio construction.**

Another indicator is that you should be able to hold the end of the drill in your hand with no discomfort immediately after you have finished the hole. These considerations are based on the assumption that most hams use the average carbon drill, and not one of the more expensive type designed to operate at high temperatures. Most tool steel will start to lose its hardness at a little over 100° C. At 600°, it is as soft as mild steel, and must be heat treated and tempered again. That means that most hams would have to grind the softened portion off and then attempt to regrind the cutting edges.

Brass should always be drilled with no lubricant. For one thing, the brass slides readily against steel. Bronze is in the same class. Witness the large number of bronze bearings in current use. Almost all of the zinc alloys may so be treated. If a lubricant is used, it usually only makes the particles cling together and thus clog up the drill point. Aluminum and its alloys are sometimes lubricated with kerosene or milk.

The drill speed (number of revolutions per minute of the drill) and the drilling feed (rate at which the drill is pushed into the work) are interdependent. The safe, simple way to determine them is to watch the temperature. If the drill is running too hot, decrease the feed. If it still runs too hot, decrease the speed. In drilling, it is a safe practice never to feed the drill in a distance greater than the diameter of the drill without backing it off until the work is clear. This permits you to examine the point, and permits the drill to clear itself of particles which may be clogging it at the point of cutting. This looks like a waste of time, but actually will be a time saver. You won't have to stop to replace broken and softened drills.

The desired speed for drilling depends on size of hole, kind of stock, rate of feed, etc. A typical job for a ham would be a lot of no. 27 or no. 19 holes in a steel chassis. The second speed on most drills, or about 1200 r.p.m., is about right. For electralloy or aluminum a step faster might be used. The 2400 r.p.m. pulley can be used for drills like no. 36 and smaller. First speed, usually about 600 r.p.m., will make those $\frac{3}{8}$ inch and $\frac{1}{2}$ inch holes. There are tables available about cutting speeds in feet per minute, etc., but experience can best tell you if the speed is right. The speed must match the feed—cut a continuous chip—and the drill should not run hot.

Plastics. One can get excellent r.f. insulation in various forms and prepare it with only a little more trouble than our old standby,

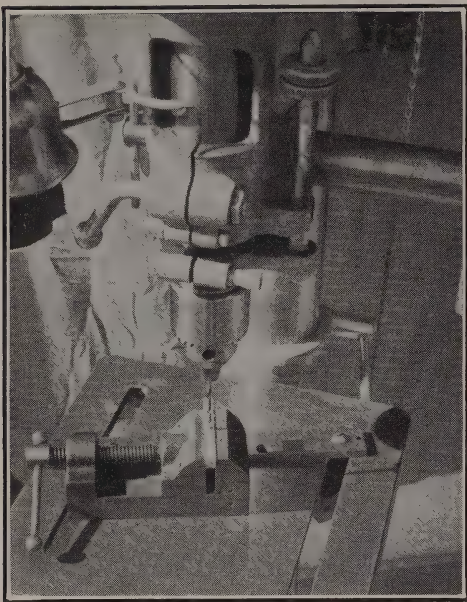


Figure 5.
DRILL PRESS VISE.
Small objects are best handled on a drill press by means of a drill press vise.

bakelite. There are two kinds of Mycalex. The kind on Cardwell condensers is known as G.E. 1364; then, there is a softer kind called leadless, which drills easier, but is more apt to chip and crack. The G.E.1364 kind can be bought in strips of various sizes, and the strips cut up with a hack saw. Sawing a wide piece would be laborious.

Drilling a piece of Mycalex is like drilling a piece of stone, and drills are bound to dull rapidly. There are special drills known as Fosdick drills that can be used to advantage if any great amount is to be drilled. Ordinary twist drills will do if they are sharpened after every hole or two. The powder which results from drilling must be removed by blowing, as fast as it forms. Another way to remove the powder is to drill the piece submerged in water. The powder will float to the top and not clog the drill. Just to lubricate the drill with water or oil in the usual way will make things worse, however. Then the powder will form a paste around the drill. Better to drill dry, with slow speed and frequent stops for cleaning the drill and hole. When the drill point starts to break through, the work should be turned over and finished on the reverse side to prevent chipping.

Lucite and polystyrene products like "Victron," "912-B," etc., drill and tap as easily as bakelite except for their notorious susceptibility to heat. Drilling speeds can be about the same as for brass or bakelite; slower if heating results. When drilling through a thick piece of the transparent kind, you can see the side wall of the hole turn white and flaky if the point warms up. A little of this roughness isn't objectionable, but keep the point cool. Drills must be sharp, and the flutes kept clear of chips. Frequent sharpening is not necessary, as these plastics are very easy on tools. If any quantity is to be worked, special "bakelite drills" with coarse flutes can be used. Some of these materials are more flexible than others; but it isn't wise to hit the center punch too hard. Use soap and water to wash the work after handling.

Danger. Most drill presses are equipped with some means of clamping the work. It is always wise to use these, unless the piece is large and the holes are small. A piece, especially of sheet steel, which gets jammed on the drill and tears out of the operator's hands, is a dangerous weapon. With small pieces, it is best to clamp them in a tool maker's vise.

When working with sheet steel (as you usually are on chassis construction), if the piece is large, you may hold it safely by hand. Wear gloves and hold the work *firmly* with both hands. The drill feed may be easily arranged to operate by foot for these operations.

Steel parallels are indispensable when the piece to be drilled is irregular. Under a panel or chassis, parallels have advantages over a block. First, they are more accurate. Then, as the work progresses, they can be moved about so as to miss the burrs that accumulate on the under side. Work that is laid flat on a block or table often gets tilted because of such burrs. Real parallels are expensive, but there are cheaper substitutes: pieces of cold-rolled steel bar and printers' "iron furniture" make excellent parallels.

Workbench Construction. While a well-built table can be used for a workbench, the greater convenience and strength obtainable in a bench designed specially for radio work makes the time and effort expended in its construction worthwhile. The bench may be of the open type,—that is, similar to a table,—but heavier and provided with a tool panel on the back, or it can be of the cabinet type in which drawers and shelves are provided in addition to the tool panel. A smooth flat top is important, the 2-inch laminated type being about the best; but one composed of $\frac{1}{8}$-inch or $\frac{1}{4}$-inch "Presdwood" glued to a sheet of $\frac{3}{4}$-inch or $\frac{7}{8}$-inch five- or seven-ply panel is very satisfactory. A good backboard can be made of two or three school-type drawing boards, fastened together with a long batten and held to the bench with large size common shelf angles.

Suitable plans are available from some tool manufacturers and home workshop magazines.

Construction Practice

Chassis Layout. The chassis first should be covered with a layer of wrapping paper, which is drawn tightly down on all sides and fastened with scotch tape. This allows any number of measurement lines and hole centers to be spotted in the correct positions without making any marks on the chassis itself. Place on it the parts to be mounted and play a game of chess with them, trying different arrangements until all the grid and plate leads are made as short as possible, tubes are clear of coil fields, r.f. chokes are in safe positions, etc. Remember, especially if you are going to use a panel, that a good mechanical layout often can accompany sound electrical design, but that the electrical design should be given first consideration.

All too often parts are grouped to give a symmetrical panel, irrespective of the arrangement behind. When a satisfactory arrangement has been reached, the mounting holes may be marked. The same procedure now must be followed for the underside, always being careful to see that there are no clashes between the two (that no top mounting screws come down into the middle of a paper condenser on the underside, that the variable condenser rotors do not hit anything when turned, etc.).

When all the holes have been spotted, they should be center-punched *through* the paper into the chassis. Don't forget to spot holes for leads which must also come through the chassis.

For transformers which have lugs on the bottoms, the clearance holes may be spotted by pressing the transformer on a piece of paper to obtain impressions, which may then be transferred to the chassis.

Punching. In cutting socket holes, one can use either a fly-cutter or socket punches. These punches are easy to operate and only a few precautions are necessary. The guide pin should fit snugly in the guide hole. This increases the accuracy of location of the socket. If this is not of great importance, one may well use a drill of $\frac{1}{32}$ inch larger diameter than the guide pin. Some of the punches will operate without guide holes, but the latter always make the punching operation simpler and easier. The only other precaution is to be sure the work is properly lined up before applying the hammer. If this is not done, the punch may slide sideways when you strike and thus not only shear the chassis but also take off part of the die. This is easily avoided by always making sure that the piece is parallel to the faces of the punch, the die, and the base. The latter should be an anvil or other solid base of heavy material.

A punch by *Greenlee* forces socket holes through the chassis by means of a screw turned with a wrench. It is noiseless, and works much more easily and accurately than most others. It requires the use of a $\frac{3}{8}$-inch center hole to accommodate the screw.

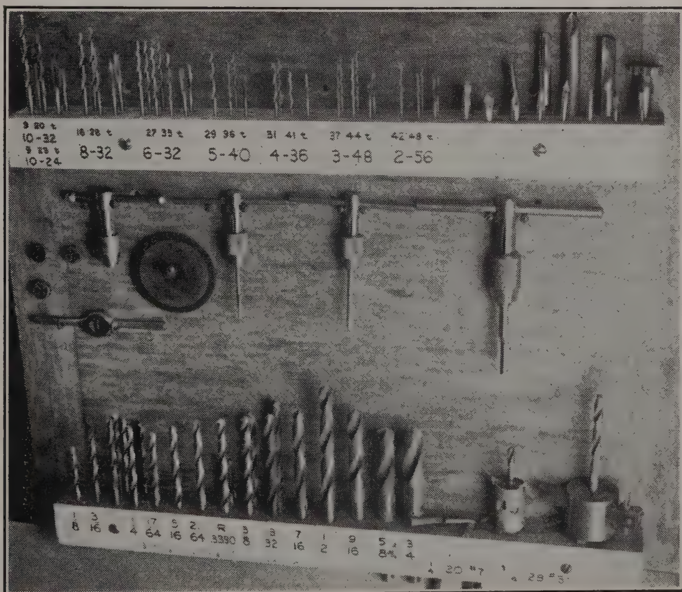


Figure 6.
TOOL MOUNTING BOARD.

An excellent assortment and mounting arrangement for power cutting tools. Note that there is a clearance drill, a tap drill, and a tap, each in its proper place, for each of the common sizes of machine screws.

Figure 7.
ENCLOSED-TYPE WORKBENCH.
An enclosed-type workbench with three drawing boards along the rear making up the backboard. The shelf along the backboard serves to strengthen it, and furnishes additional storage space.

Transformer Cutouts. Cutouts for transformers and chokes are not so simply handled. After marking off the part to be cut, drill about a $\frac{1}{4}$-inch hole on each of the inside corners and tangential to the edges. After burring the holes, clamp the piece and a block of cast iron or steel in the vise. Then, take your burring chisel and insert it in one of the corner holes. Cut out the metal by hitting the chisel with a hammer. The blows should be light and numerous. The chisel acts against the block in the same way that the two blades of a pair of scissors work against each other. This same process is repeated for the other sides. A file is used to trim up the completed cutout.

Another method is to drill the four corner holes large enough to take a hack saw blade, then saw instead of chisel. The four holes permit nice looking corners.

Removing Burrs. In both drilling and punching, a burr is usually left on the work. There are three simple ways of removing these. Perhaps the best is to take a chisel (be sure it is one for use on metal) and set it so that its bottom face is parallel to the piece. Then gently tap it with a hammer. This usually will make a clean job with a little practice. If one has access to a counterbore, this will also do a nice job. A countersink will work, although it bevels the edges. A drill of several sizes larger is a much used arrangement. The third method is by filing off the burr, which does a good job but scratches the adjacent metal surfaces badly.

Mounting Components. There are two methods in general use for the fastening of transformers, chokes, and similar pieces of apparatus to chassis or breadboards. The first, using nuts and machine screws, is slow, and the commercial manufacturing practice of using self-tapping screws is gaining favor. For the mounting of small parts such as resistors and condensers, tie strips are very useful to gain rigidity. They also contribute materially to the appearance of finished apparatus.

Grommets of the proper size, placed in all chassis holes through which wires are to be passed, will give a neater appearing job and also will reduce the possibility of short circuits.

Soldering. Making a strong, low-resistance solder joint does not mean just dropping a blob of solder on the two parts to be joined and then hoping that they'll stick. There are several definite rules that *must* be observed.

All parts to be soldered must be absolutely clean. To clean a wire, lug, or whatever it may be, take your pocket knife and scrape it thoroughly, until fresh metal is laid bare. It is not enough to make a few streaks; scrape until the part to be soldered is bright.

Make a good mechanical joint before applying any solder. Solder is intended primarily to make a good *electrical* connection; mechanical rigidity should be obtained by bending the wire into a small hook at the end and nipping it firmly around the other part, so that it will hold well even before the solder is applied.

Keep your iron properly tinned. It is impossible to get the work hot enough to take the solder properly if the iron is dirty. To tin your iron, file it, while hot, on one side until a full surface of clean metal is exposed. Immediately apply rosin core solder until a thin layer flows completely over the exposed surface. Repeat for the other faces. Then take a clean rag and wipe off all excess solder and rosin. The iron should also be wiped frequently while the actual construction is going on; it helps prevent pitting the tip.

Apply the solder to the work, not to the iron. The iron should be held against parts to be joined until they are thoroughly heated. The solder should then be applied against the parts, and the iron should be held in place until the solder flows smoothly and envelops the work. If it acts like water on a greasy plate, and forms a ball, the work is not sufficiently clean.

The completed joint must be held perfectly still until the solder has had time to solidify. If the work is moved before the solder has become *completely* solid, a "cold" joint will result. This can be identified immediately, because the solder will have a dull "white" appearance rather than one of shiny "silver." Such joints tend to be of high resistance, and will very likely have a bad effect upon a circuit. The cure is simple: merely reheat the joint and do the job correctly.

Wipe away all surplus flux when the joint has cooled if you are using a paste type flux. Be sure it is non-corrosive, and use it with plain (not rosin core) solder.

Finishes. If the apparatus is constructed on a painted chassis (commonly available in black crackle and gray crackle), there is no need for application of a protective coating when the equipment is finished, assuming that you are careful not to scratch or mar the finish while drilling holes and mounting parts. However, many amateurs prefer to use unpainted (zinc or cadmium plated) chassis, because it is much simpler to make a chassis ground connection with this type of chassis. A thin coat of clear "linoleum" lacquer may be applied to the whole chassis after the wiring is completed to retard rusting. In localities near the sea coast it is a good idea to lacquer the various chassis cut-outs even on a painted chassis, as rust will get a good start at these points unless the metal is protected where the drill or saw has exposed it. If too thick a coat is applied, the lacquer will tend to peel. It may be thinned with lacquer thinner to permit application of a light coat. A thin coat will adhere to any *clean* metal surface that is not too shiny.

An attractive dull gloss finish, almost velvety, can be put on aluminum by sand-blasting it with a very weak blast and fine particles and then lacquering it. Soaking the aluminum in a solution of lye produces somewhat the same effect as a fine grain sand blast.

There are also several brands of dull gloss black enamels on the market which adhere well to metals and make a nice appearance. Air-drying crackle finishes are sometimes successful, but a baked job is usually far better. Crackle finishes, properly applied, are very durable and are pleasing to the eye. If you live in a large community, there is probably an enamelling concern which can crackle your work for you at a reasonable cost. A very attractive finish, for panels especially, is to spray a crackle finish with aluminum paint. In any painting operation (or plating, either, for that matter), the work should be very thoroughly cleaned of all greases and oils.

To protect brass from tarnish, thoroughly cleanse and remove the last trace of grease by the use of potash and water. The brass must be carefully rinsed with water and dried; but in doing it, care must be taken not to handle any portion with the bare hands or anything else that is greasy. Then lacquer.

Drilling Glass. This is done very readily with a common drill by using a mixture of turpentine and camphor. When the point of the drill has come through, it should be taken out and the hole worked through with the point of a three-cornered file, having the edges ground sharp. Use the corners of the file, scraping the glass rather than using the file as a reamer. Great care must be taken not to crack the glass or flake off parts of it in finishing the hole after the point of the drill has come through. Use the mixture freely during the drilling and scraping. The above mixture will be found useful in drilling hard cast iron.

Etching Solution. Add three parts nitric acid to one part muriatic acid. Cover the piece to be etched with beeswax. This can be done by heating the piece in a gas or alcohol flame and rubbing the wax over the surface. Use a sharp steel point or hard lead pencil point as a stylus. A pointed glass dropper can be used to put the solution at the place needed. After the solution foams for two or three minutes, remove with blotting paper and put oil on the piece, and then heat and remove the wax.

Chromium Polish. So much chromium is now used in radio sets and on panels that it is well to know that this finish may be polished. The only materials required are absorbent cotton or soft cloth, alcohol, and ordinary lampblack.

A wad of cotton or the cloth is moistened in the alcohol and pressed into the lampblack. The chromium is then polished by rubbing the lampblack adhering to the cotton briskly over its surface. The mixture dries almost instantly and may be wiped off with another wad of cotton.

The alcohol serves merely to moisten the lampblack to a paste and make it stick to the cotton. The mixture cleans and polishes very quickly and cannot scratch the chromium surface. It polishes nickel-work just as effectively as it does chromium. Care should be taken to see that the lampblack does not contain any hard, gritty particles which might produce scratches during the polishing.

Broadcast Interference

RADIO signals which intrude upon a broadcast program constitute a nuisance to which disturbed listeners are bound to object vigorously.

Broadcast interference is a matter of grave importance to all amateurs. Indeed, an amateur station license is placed in considerable jeopardy by repeated citations of interference with broadcast or other commercial stations. The FCC regulations are particularly severe in this respect, and they require that the offending amateur correct the trouble or keep off the air during specified hours of the day or night.

In general, signals from a transmitter operating properly are not picked up by receivers tuned to other frequencies unless the receiver is of inferior design, or is in poor condition. Therefore, if the receiver is of good design and is in good repair, the burden of rectifying the trouble rests with the owner of the interfering station.

Phone and c.w. stations both are capable of causing broadcast interference, key-click annoyance from code transmitters being particularly objectionable. The elimination of key clicks is fully covered in Chapter 7 under *Keying*.

A knowledge of each of the several types of broadcast interference, their cause, and methods of eliminating them is necessary to the successful disposition of this trouble. An effective method of combatting one variety of interference is often of no value whatever in the correction of another type. Broadcast interference seldom can be cured by "rule of thumb" procedure.

Broadcast interference, as covered in this chapter, refers to standard (amplitude modulated, 550–1600 kc.) broadcast. Interference to frequency modulated broadcast is highly unlikely except when there is an f.m. receiver in close proximity to a transmitter afflicted with h.f. parasitics or radiating strong harmonics. As such radiation is illegal, it is assumed that no such interference is experienced except in rare instances. When it does occur, it calls for suppression of parasitics or harmonics at the transmitter, a subject covered under *Transmitter Theory* and in the *Antenna* chapter.

Interference Classifications

Depending upon whether it is traceable directly to causes within the *station* or within the *receiver*, broadcast interference may be divided into two main classes. For example, that type of interference due to transmitter overmodulation is at once listed as being caused by improper operation, while an interfering signal that tunes in and out with a broadcast station is probably an indication of cross-talk in the receiver, and the poorly-designed input stage of the receiver is held liable. The various types of interference and recommended cures will be discussed separately.

Blanketing. This is not a tunable effect, but a total blocking of the receiver. A more or less complete "washout" covers the entire receiver range when the carrier is switched on. This produces either a complete blotting out of all broadcast stations, or else knocks down their volume several decibels—depending upon the severity of the interference. Voice modulation of the carrier causing the blanketing will be highly distorted or even unintelligible. Keying of the carrier which produces the blanketing will cause an annoying fluctuation in the volume of the broadcast signals.

Blanketing generally occurs in the immediate neighborhood (inductive field) of a powerful transmitter, the affected area being directly proportional to the power of the transmitter. This type of interference occurs most frequently where the receiver uses an outside antenna which happens to resonate at a frequency close to that of the offending transmitter. Also it is more prevalent with transmitters which operate in the 80- and 160-meter bands, than with those on the higher frequencies.

The remedies are to (1) shorten the receiving antenna, and thereby shift its resonant frequency, or (2) remove it to the interior of the building, (3) change the direction of either the

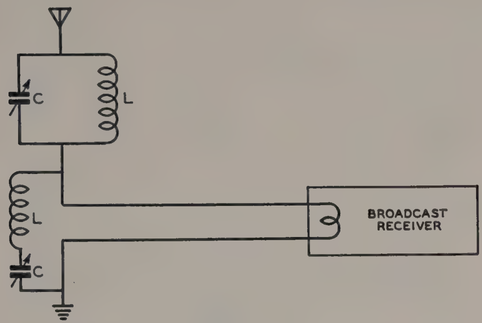


Figure 1.
EFFECTIVE WAVE TRAP CIRCUIT FOR HIGH ATTENUATION OF INTERFERING SIGNAL REACHING RECEIVER VIA ANTENNA.

This type of trap works at full efficiency over but a small range in frequency, and therefore is not effective when several interfering signals of widely different frequencies are present. When only moderate attenuation is required, a single tank (either series or shunt) will often suffice. For coil and condenser values refer to Figure 3.

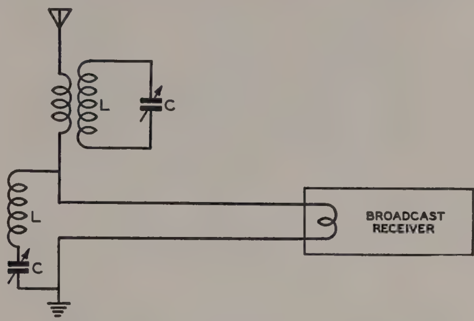


Figure 2.
MODIFICATION OF CIRCUIT SHOWN IN FIGURE 1.

In this case, the parallel resonant tank is coupled to the antenna with 3 to 6 turns of wire instead of being placed in series with the antenna lead. It gives slightly better performance than the circuit of Figure 1 with certain antennas.

receiving or transmitting antenna to minimize their mutual coupling, or (4) keep the interfering signal from entering the receiver input circuit by installing a wave-trap tuned to the signal frequency (see Figure 1).

A suitable wave-trap is quite simple in construction, consisting only of a coil and midget variable condenser. When the trap circuit is tuned to the frequency of the interfering signal, little of the interfering voltage reaches the grid of the first tube.

The wave-trap must be installed as close to the receiver antenna terminal as practicable, hence it should be as small in size as possible. The variable condenser may be a midget air-tuned trimmer type, and the coil may be wound on a 1-inch dia. form. The table of Figure 3 gives winding data for wave-traps built around a 50-$\mu\mu$fd. variable condenser. For best results, both a shunt and a series trap should be employed as shown.

Figure 2 shows a two-circuit coupled wave-trap that is somewhat sharper in tuning and more efficacious. The specifications for the coil L_2 may be obtained from the table in Figure 3. The primary, L_1, consists of 3 to 5 closewound turns of the same size wire wound in the same direction on the same form as L_2, and separated from the latter by $\frac{1}{8}$ of an inch.

Figure 3. R. F. WAVE TRAP COIL AND CONDENSER TABLE.

BAND	COIL L	CONDENSER C
160	41 turns No. 28 enameled close-wound 1-inch form	50-$\mu\mu$fd. variable shunted by 200-$\mu\mu$fd. fixed mica
80	41 turns No. 28 enameled close-wound 1-inch form	50-$\mu\mu$fd. variable
40	21 turns No. 24 enameled 11/16-inch long 1-inch form	50-$\mu\mu$fd. variable
20	7 turns No. 24 enameled 5/16-inch long 1-inch form	50-$\mu\mu$fd. variable
10	4 turns No. 24 enameled 5/16-inch long 1-inch form	50-$\mu\mu$fd. variable

Overmodulation. A carrier modulated in excess of 100 per cent acquires sharp cutoff periods (Figure 4) which give rise to high damping. This creates a broad signal and often generates spurious frequencies at odd places on the dial. High damping of a radiotelephone signal may at the same time bring about impact or shock excitation of nearby receiving antenna and power lines, transmitting interfering voltages in that manner.

Broadcast interference due to overmodulation is generally common to 160- and 75-meter operation. The remedy is to reduce the modulation percentage.

Cross Modulation. Cross modulation or "cross talk" is characterized by the amateur signal "riding in" on top of strong local broad-

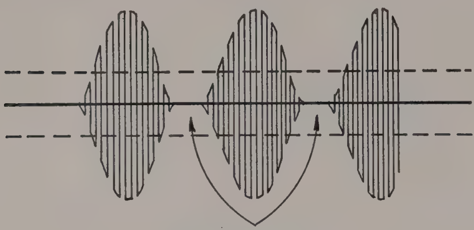


Figure 4.
ILLUSTRATING HIGH DAMPING CHARACTERISTIC OF BADLY OVER-MODULATED SIGNAL.
The resulting interference seldom can be cured by wave traps or line filters; it must be corrected at the transmitter.

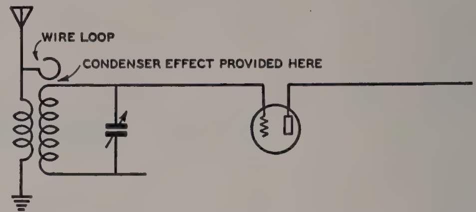


Figure 5.
TYPICAL AUXILIARY CAPACITY COUPLING CIRCUIT USED IN B.C. SETS TO BOOST GAIN AT 1500 KC. END OF BAND.
Even though the coupling capacity may be small, it will have a fairly low reactance at high frequencies, and will aggravate interference from amateur stations, particularly those working on 14 and 28 Mc.

casts. There is usually no heterodyne note, the amateur signal being tuned in and out with the program carriers.

This effect is due entirely to a faulty input stage in the affected receiver. Modulation of the interfering carrier will swing the operating point of the input tube. This type of trouble is seldom experienced when a variable-μ tube is used in the input stage.

Where the receiver is too ancient to incorporate such a tube, and is probably poorly shielded at the same time, it will be better to attach a wave-trap of the type shown in Figure 1 than to attempt rebuilding of the receiver. The addition of a good ground and a shield can over the input tube often adds to the effectiveness of the wave-trap.

Transmission via Capacity Coupling. A small amount of capacity coupling is now widely used in receiver r.f. and detector transformers as a gain booster at the high-frequency end of the tuning range. The coupling capacity is obtained by means of a small loop of wire cemented close to the grid end of the secondary winding, and with one end directly connected to the plate or antenna end of the primary winding (see Figure 5).

From the relations of capacitive reactance, it is easily seen that a small condenser will favor the higher frequencies, and it is evident that capacity coupling in the receiver coils will tend to pass amateur short-wave signals into a receiver tuned to broadcast frequencies.

The amount of capacity coupling may be reduced to eliminate interference by moving the coupling turn farther away from the secondary coil. However, a simple wave-trap of the type shown in Figures 1 and 2, inserted at the antenna input terminal, will generally accomplish the same result and is more to be recommended than changing the capacity coupling (which lowers the receiver gain at the high frequency end of the broadcast band). Should the wave-trap alone not suffice, it will be necessary to resort to a reduction in capacity coupling.

In some simple broadcast receivers, capacity coupling is unintentionally obtained by too closely coupled primary and secondary coils, or as a result of running a long primary or antenna lead close to the secondary coil of an unshielded antenna coupler.

Phantoms. When two strong local carriers are separated by a certain number of kilocycles, the beat note resulting between them may fall on some frequency within the broadcast band and, if rectified by any means, be audible at that point. If such a "phantom" signal falls on a local broadcast frequency, there will be heterodyne interference as well. This is a common occurrence with broadcast receivers in the neighborhood of two amateur stations, or an amateur and a police station. It also sometimes occurs when only one of the stations is located in the immediate vicinity.

As examples: the beat note between amateur carriers on 2000 kc. and 3500 kc. falls on 1500 kc., and an 1812-kc. amateur signal might beat with a local 1712-kc. police carrier to produce a 100-kc. phantom. And, if the latter two carriers are strong enough, harmonics will be encountered every 100 kilocycles throughout the broadcast band,—that is, if rectification of the signals takes place anywhere in the vicinity. A poor contact between two oxidized wires can produce rectification.

Two stations must be transmitting simultaneously to produce a phantom signal; when either station goes off the air the phantom disappears. Hence, this type of interference is apt to be reported as highly intermittent, and might be difficult to duplicate unless a test

oscillator is used "on location" to simulate the missing station. Such interference cannot be remedied at the transmitter, and often the rectification takes place some distance from the receivers. In such occurrences it is most difficult to locate the source of the trouble.

It will also be apparent that a phantom might fall on the intermediate frequency of a simple superhet receiver and cause interference of the untunable variety if the manufacturer has not provided an i.f. wave-trap in the antenna circuit. Examples of this occurrence are the 175-kc. beat between 1887 (amateur) and 1712 kc. (police) or between amateur signals on 1820 kc. and 1995 kc.

This particular type of phantom may, in addition to causing i.f. interference, generate harmonics which may be tuned in and out with heterodyne whistles from one end of the receiver dial to the other. It is in this manner that "birdies" often result from the operation of nearby amateur stations.

When one component of a phantom is a steady, unmodulated carrier, only the intelligence present on the other carrier is conveyed to the broadcast receiver.

Phantom signals almost always may be identified by the suddenness with which they are interrupted, signalizing withdrawal of one party to the union. This is especially baffling to the inexperienced interference-locater, who observes that the interference suddenly disappears, even though his own transmitter remains in operation.

If the mixing or rectification is taking place in the receiver itself, a phantom signal may be eliminated by removing either one of the contributing signals from the receiver input circuit. A wave-trap of the types shown in Figures 1 and 2, tuned to either signal, will do the trick. If the rectification is taking place outside the receiver, the wave-trap should be tuned to the frequency of the phantom, instead of to one of its components. I.f. wave-traps may be built around a 2.5-millihenry r.f. choke as the inductor, and a compression-type mica padding condenser. The condenser should have a capacity range of 250–525 $\mu\mu$fd. for the 175- and 206-kc. intermediate frequencies; 65–175 $\mu\mu$fd. for 260 kc. and other intermediates lying between 250 and 400 kc.; and 17–80 $\mu\mu$fd. for 456, 465, 495, and 500 kc. Slightly more capacity will be required for resonance with a 2.1 millihenry choke.

Spurious Emissions. This sort of interference arises from the transmitter itself. The radiation of any signal (other than the intended carrier frequency) by an amateur station is prohibited by FCC regulations. Spurious radiation may be traced to imperfect neutralization, parasitic oscillations in the r.f. or modulator stages, or to "broadcast-band" v.f. oscillators.

Low-frequency parasitics may actually occur on broadcast frequencies or their near subharmonics, causing direct interference to programs. An all-wave monitor operated in the vicinity of the transmitter will detect these spurious signals.

The remedy will be obvious in individual cases. Elsewhere in this book are discussed methods of complete neutralization and the suppression of parasitic oscillations in r.f. and audio stages.

Stray Receiver Rectification. A receiver in the immediate neighborhood of a strong transmitter is subject to stray rectification within the receiver. It is due to the interfering signal being rectified by the second detector in a superhet (detector in a tuned r.f. set), or an audio stage of the receiver if poorly shielded or containing too long a grid lead.

This type of interference is most commonly caused by ultra-high-frequency transmitters, doubtless because at those frequencies lengthy connections in the receiver can easily become fractions of the transmitter wavelength. The interfering signal is not tunable, and generally covers the entire dial.

If the receiver is not a series-filament set, the trouble may be localized by removing the tubes, starting with the input stage and working toward the audio output stage. The interfering signal will cease when the tube rectifying it is removed from its socket.

Signal rectification in an audio stage may be cured by connecting a 2.5-millihenry pi-wound r.f. choke in series with the control-grid lead and input terminal and a .0001 μfd. condenser from grid to ground. But the task is not so simple when rectification occurs in one of the other stages. Here, complete shielding of the set, tubes, and exposed r.f. leads (such as top-cap grid leads) will have to be provided. In addition, it may be necessary to lower the bias of the offending stage.

"Floating" Volume Control Shafts. Several sets have been encountered where there was only a slightly interfering signal; but, upon placing one's hand up to the volume control, the signal would greatly increase. Investigation revealed that the volume control was installed with its shaft insulated from ground. The control itself was connected to a critical part of a circuit, in many instances to the grid of a high-gain audio stage. The cure is to install a volume control with *all* the terminals insulated from the shaft, and then to ground the shaft.

Spray-Shield Tubes. Although they are no longer made, there are yet quite a few sets in use which employ spray-shield tubes. These

BAND	COIL L	CONDENSER C
160	26 turns No. 14 enameled 4-inch diameter 3-inch length	100-$\mu\mu$fd. variable
80	17 turns No. 14 enameled 3-inch diameter 2¼-inch length	100-$\mu\mu$fd. variable
40	11 turns No. 14 enameled 2½-inch diameter 1½-inch length	100-$\mu\mu$fd. variable
20	4 turns No. 10 enameled 3-inch diameter 1⅛-inch length	100-$\mu\mu$fd. variable
10	3 turns ¼-inch o.d. copper tubing 2-inch diameter 1-inch length	100-$\mu\mu$fd. variable

Figure 6. POWER-LINE WAVE TRAP COIL AND CONDENSER TABLE.

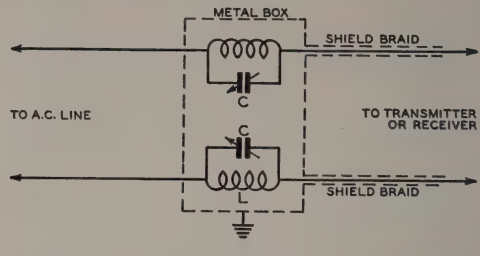


Figure 7.
METHOD OF CONNECTING POWER LINE WAVE TRAP.
A parallel resonant circuit is more effective than an r.f. choke in keeping r.f. from getting from a transmitter into the power line, or from the power line into a receiver. A .05-μfd. tubular condenser connected from each 110-volt wire to ground often will increase the effectiveness of the traps. They may be connected on either side of the line traps.

are used in both r.f. and in audio circuits. In some audio applications of this type of tube, the cathode and the spray-shield (to which the cathode is connected) are not at ground potential, but are by-passed to ground with an electrolytic condenser of large capacity. This type of condenser is a very poor r.f. filter and, in a strong r.f. field, some detection will take place, producing interference. The best cure is to install a standard glass tube with a glove shield, which is then actually grounded, and also to shield the grid leads to these tubes. As an alternative, bypassing the electrolytic cathode condenser with a .05 μfd. tubular paper condenser may be tried.

Power-Line Pickup. When radio-frequency energy from a radio transmitter enters a broadcast receiver through the a.c. power lines, it has either been fed back into the lighting system by the offending transmitter, or picked up from the air by overhead power lines. Underground lines are seldom responsible for spreading this form of interference.

To check the path whereby the interfering signals reach the lines, it is only necessary to replace the transmitting antenna with a dummy antenna and adjust the transmitter for maximum output. If the interference then ceases, overhead lines have been picking up the energy. The trouble can be cleared up only by installing wave-traps in the power lines at the receiver. These are then tuned to the interfering signal frequency. If the receiver is reasonably close to the transmitter, it is very doubtful that changing the direction of the transmitting antenna to right angles with the overhead lines will eliminate the trouble.

If, on the contrary, the interference continues when the transmitter is connected to the dummy antenna, radio-frequency energy is being fed directly into the power line by the transmitter, and the station must be inspected to determine the cause.

One of the following reasons for the trouble will usually be found: (1) the r.f. stages are not sufficiently bypassed and/or choked, (2) the antenna coupling system is not performing efficiently, (3) the power transformers have no electrostatic shields; or, if shields are present, they are ungrounded, (4) power lines are running too close to an antenna or r.f. circuits carrying high currents. If none of these causes apply, wave-traps must be installed in the power lines at the transmitter to remove r.f. energy passing back into the lighting system.

The wave-traps used in the power lines at transmitter or receiver must be capable of passing relatively high amperage. The coils are accordingly wound with heavy wire. Figure 6 lists the specifications for power line wave-trap coils, while Figure 7 illustrates the method of connecting these wave-traps. Observe that these traps are enclosed in a shield box of heavy iron or steel, well grounded.

All-Wave Receivers. Each complete-coverage home receiver is a potential source of annoyance to the transmitting amateur. The novice short-wave broadcast-listener who tunes in an amateur station often considers it an interfering signal, and complains accordingly.

Neither selectivity nor image rejectivity in most of these sets is in any wise comparable

to those properties in a communication receiver. The result is that an amateur signal will occupy too much dial space and appear at more than one point, giving rise to interference on adjacent channels and removed channels as well.

If carrier-frequency harmonics are present in the amateur transmission, serious interference will result at the all-wave receiver. The harmonics will, if the carrier frequency has been so unfortunately chosen, fall directly upon a favorite short-wave broadcast station and arouse warranted objection.

The amateur is apt to be blamed, too, for transmissions for which he is not responsible, so great is the public ignorance of short-wave allocations and signals. Owners of all-wave receivers have been quick to ascribe to amateur stations all signals they hear from tape machines and V-wheels, as well as stray tones and heterodyne flutters they hear.

The amateur cannot be held responsible when his carrier is deliberately tuned in on an all-wave receiver. Neither is he accountable for the width of his signal on the receiver dial, or for the strength of image repeat points, if it can be proven that the receiver design does not afford good selectivity and image rejection.

If he so desires, the amateur (or the owner of the receiver) might sharpen up the received signal somewhat by shortening the receiving antenna. Set retailers often supply quite a sizable antenna with all-wave receivers, but most of the time these sets perform almost as well with a few feet of inside antenna.

The amateur *is* accountable for harmonics of his carrier frequency. Such emissions are unlawful in the first place, and he must take all steps necessary to their suppression. Practical suggestions for the elimination of harmonics will be found elsewhere in this book (see *Index*).

Superheterodyne Interference

In addition to those types of interference already discussed, there are two more which are common to superhet receivers. The prevalence of these types is of great concern to the amateur, although the responsibility for their existence more properly rests with the broadcast receiver.

The first is the production of broadcast-band images by 160-meter amateur stations. This is possible since the separation between the broadcast band and the 160-meter region is small enough to establish image-frequency relationships.

The mechanism whereby image production is accomplished may be explained in the following manner: when the first detector is set to the frequency of an incoming signal, the high-frequency oscillator is operating on another frequency which differs from the signal by the number of kilocycles in the intermediate frequency. Now, with the setting of these two stages undisturbed, there is another signal which will beat with the high-frequency oscillator to produce an i.f. voltage. This other signal is the so-called image, which is separated from the desired signal by twice the intermediate frequency.

Thus, in a receiver with 175-kc. i.f., tuned to 1000 kc.: the h.f. oscillator is operating on 1175 kc., and a signal on 1350 kc. (1000 kc. plus 2×175 kc.) will beat with this 1175 kc. oscillator frequency to produce the 175-kc. i.f. signal. Similarly, when the same receiver is tuned to 1400 kc., an amateur signal on 1750 kc. can come through. The dial point where any 160-meter signal will produce an image can be determined from the equation:

$$F = (F_{am} - 2\text{ i.f.})$$

Where F_b = receiver dial frequency,
F_{am} = amateur transmitter frequency, and
i.f. = receiver intermediate frequency.

If the image appears only a few cycles or kilocycles from a broadcast carrier, heterodyne interference will be present as well. Otherwise, it will be tuned in and out in the manner of a station operating in the broadcast band. Sharpness of tuning will be comparable to that of broadcast stations producing the same a.v.c. voltage at the receiver.

The second variety of superhet interference is the result of harmonics of the receiver h.f. oscillator beating with amateur carriers to produce the intermediate frequency of the receiver. The amateur transmitter will always be found to be on a frequency equal to some harmonic of the receiver h.f. oscillator, *plus or minus the intermediate frequency.*

As an example: when a broadcast superhet with 465-kc. i.f. is tuned to 1000 kc., its high-frequency oscillator operates on 1465 kc. The third harmonic of this oscillator frequency is 4395 kc., which will beat with an amateur 'phone signal on 3950 kc. to send a signal through the i.f. amplifier. The 3950 kc. signal would be tuned in at the 1000-kc. point on the dial.

Some oscillator harmonics are so related to amateur frequencies that more than one point of interference will occur on the receiver dial. Thus, a 3500-kc. signal may be tuned in at six points on the dial of a nearby broadcast superhet having 175 kc. i.f. and no r.f. stage.

Insofar as remedies for image and harmonic superhet interference are concerned, it is well

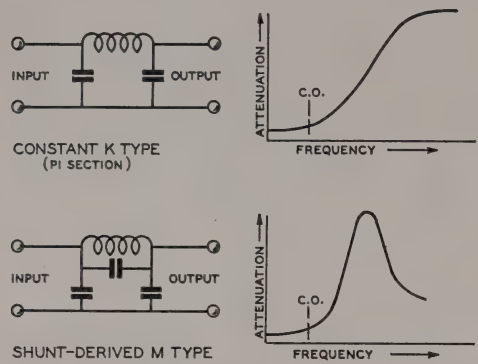


Figure 8.

TWO TYPES OF LOW PASS FILTERS AND THE KIND OF ATTENUATION CURVE OBTAINED WITH EACH.

The M-derived type has sharper cut-off but not as great attenuation at frequencies two or more octaves above the cut-off frequency.

to remember that *if* the amateur signal did not in the first place reach the input stage of the receiver, the annoyance would not have been created. It is therefore good policy to try to eliminate it by means of a wave-trap. Broadcast superhets are not always the acme of good shielding, however, and the amateur signal is apt to enter the circuit through channels other than the input circuit. If a wave-trap or filter will not cure the trouble, the only alternative will be to attempt to select a transmitter frequency such that neither image nor harmonic interference will be set up on favorite stations in the susceptible receivers. The equation given earlier may be used to determine the proper frequencies.

Low Pass Filters. The greatest drawback of the wave-trap is the fact that it is a single-frequency device; i.e.—it may be set to reject at one time only one frequency (or, at best, an extremely narrow band of frequencies). Each time the frequency of the interfering transmitter is changed, every wave-trap tuned to it must be retuned.

A much more satisfactory device is the *wave filter* which requires no tending. One type, the low pass filter, passes all frequencies below one critical frequency, and alternates all higher frequencies. It is this property that makes the device ideal for the task of removing amateur frequencies from broadcast receivers.

A good low pass filter designed for maximum attenuation at some frequency inside the

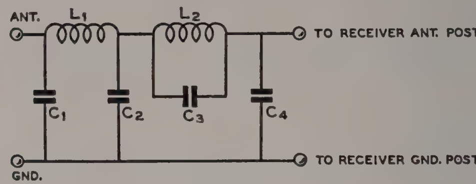


Figure 9.

COMPOSITE LOW PASS FILTER POSSESSING ADVANTAGES OF BOTH K SECTION AND M DERIVED FILTER.

This filter is highly effective in reducing broadcast interference from all high frequency stations, and requires no tuning. Constants for 400 ohm terminal impedance and 1600 kc. cut-off are as follows: L_1, 65 turns no. 22 d.c.c. close wound on $1\frac{1}{2}$ in. dia. form. L_2, 41 turns ditto, not coupled to L_1. C_1, 250 $\mu\mu$fd. fixed mica condenser. C_2, 400 $\mu\mu$fd. fixed mica condenser. C_3 and C_4, 150 $\mu\mu$fd. fixed mica condensers, former of 5% tolerance. With some receivers, better results will be obtained with a 200 ohm carbon resistor inserted between the filter and antenna post on the receiver. With other receivers the effectiveness will be improved with a 600 ohm carbon resistor placed from the antenna post to the ground post on the receiver. The filter should be placed as close to the receiver terminals as possible.

lower frequency edge of the 160-meter band will pass all broadcast carriers, but will reject signals originating in any amateur band. Naturally such a device should be installed only in standard broadcast receivers, never in all-wave sets.

Two types of low pass filters are shown in Figure 8. A composite arrangement comprising a section of each type is more effective than either type operating alone. A composite filter composed of one K-section and one shunt-derived M-section is shown in Figure 9, and is highly recommended. The M-section is designed to have maximum attenuation in the middle of the 160-meter 'phone band, and for that reason C_3 should be of the "close tolerance" variety. Likewise, C_3 should not be stuffed down inside L_2 in the interest of compactness, as this will alter the inductance of the coil appreciably, and likewise the resonant frequency.

If a fixed 150 $\mu\mu$fd. mica capacitor of 5 per cent tolerance is not available for C_3, a compression trimmer covering the range of 125–175 $\mu\mu$fd. may be substituted and adjusted to give maximum attenuation at about 1900 kc.

Radio Mathematics and Calculations

INTO this chapter have been grouped certain methods of computation and various charts which will be found useful in radio work. The decibel, logarithms, calculation of gain and loss, calculation of inductance by means of charts, and determination of reactance and of frequency of resonance will be discussed.

The Decibel

The decibel unit as used in radio engineering is actually a method of expressing a ratio between two power or energy levels. Since the ratio between two energy levels may be extremely large (several million to 1 in ordinary speech amplifier calculations), it is more convenient to make the method of expressing these ratios a logarithmic one so that the energy levels will be compacted. So it is necessary that the reader be familiar with the use of logarithms before an understanding of the decibel scale may be possible. Hence, the following few paragraphs will be devoted to an explanation of the use of logarithms.

Logarithms in Decibel Calculations. The table of logarithms presented here does not differ essentially from any other table of logarithms except that no proportional parts are given and the figures are stated to only three decimal places; this arrangement does not permit of great accuracy, but has been found to be satisfactory for all practical decibel calculations. A complete exposition on logarithms is outside the scope of this HANDBOOK; however, the very essentials together with the practical use of the tables and their application to decibels are given herewith. The following discussion is not concerned with the study of logarithms other than in their direct employment in calculating decibels gain and loss.

By definition, the logarithm of a number to the base 10 is the power to which 10 must be raised to equal that number. Hence, a logarithm to the base 10 can be considered as being an exponent of 10. Logarithms to the base 10 are called common logs. There is another type of logarithm to the base *e* (2.716) which is used in other engineering calculations. But common logarithms (base 10) are the ones used in decibel calculations.

The logarithm of a number usually consists of two parts: a whole number called the *characteristic* and a decimal called the *mantissa.* The characteristic is the integral portion to the *left* of the decimal point (see examples below), and the mantissa is the value placed to the *right.* The mantissa is all that appears in the table of logarithms.

In the logarithm, the mantissa is independent of the position of the decimal point, while the characteristic is dependent only on the position of the number with relation to the decimal point. Thus, in the following examples:

	Number		Logarithm
(a)	4021.	=	3.604
(b)	402.1	=	2.604
(c)	40.21	=	1.604
(d)	4.021	=	0.604
(e)	.4021	=	−1.604
(f)	.04021	=	−2.604

it will be seen that the characteristic is equal, algebraically, to the number of digits *minus one* to the left of the decimal point.

In (a) the characteristic is 3, in (b) 2, in (d) 0, in (e) −1, in (f) −2. The following should be remembered: (1) that for a number greater than 1, the characteristic is *one less* than the number of digits to the left of the decimal point; (2) that for a number wholly a decimal, the characteristic is *negative* and is numerically *one greater* than the number of ciphers immediately following the decimal point. Notice (e) and (f) in the above examples.

To find a common logarithm of any number, proceed as directed herewith. Suppose the number to be 5576. First, determine the

characteristic. An inspection will show that this number will be 3. The figure is placed to the *left* of a decimal point. The mantissa is now found by referring to the logarithm table. Proceed, selecting the first two numbers which are 55, then glance down the N column until coming to these figures. Advance to the right until coming in line with the column headed 7; the number will be 746. (Note that the column headed 7 corresponds to the *third* figure in the number 5576). Place the mantissa 746 to the *right* of the decimal point making the number now read 3.746. This is the logarithm of 5576. *Important:* do not consider the last figure, 6, in the number 5576 when looking for the mantissa in the accompanying three-place tables; in fact, disregard all digits beyond the first three when determining the mantissa. (*Interpolation,* to find the true log of 5576, cannot be accurately done from three-place values.) However, be doubly sure to include *all* figures when ascertaining the magnitude of the *characteristic.*

Practical application of logarithms to decibel calculations will follow. Other methods of using logarithms will be discussed as the subject develops.

Power Levels. In the design of radio devices and amplifying equipment, the standard power level of six milliwatts (.006 w.) is the arbitrary reference level of zero decibels. All power levels above the reference level are designated as plus quantities, and below as minus. The figure is always prefixed by a plus (+) or minus (—) sign indicating the direction in which the quantity is to be read.

THREE-PLACE LOGARITHMS

N	0	1	2	3	4	5	6	7	8	9
00	000	000	000	000	000	000	000	000	000	000
10	000	004	008	012	017	021	025	029	033	037
11	041	045	049	053	056	060	064	068	071	075
12	079	082	086	089	093	096	100	103	107	110
13	113	117	120	123	127	130	133	136	139	143
14	146	149	152	155	158	161	164	167	170	173
15	176	179	181	184	187	190	193	195	198	201
16	204	206	209	212	214	217	220	222	225	227
17	230	233	235	238	240	243	245	248	250	252
18	255	257	260	262	264	267	269	271	274	276
19	278	281	283	285	287	290	292	294	296	298
20	301	303	305	307	309	311	313	316	318	320
21	322	324	326	328	330	332	334	336	338	340
22	342	344	346	348	350	352	354	356	358	359
23	361	363	365	367	368	371	372	374	376	378
24	380	382	383	385	387	389	390	392	394	396
25	397	399	401	403	404	406	408	409	411	413
26	415	416	418	420	421	423	424	426	428	429
27	431	433	434	436	437	439	440	442	444	445
28	447	448	450	451	453	454	456	457	459	460
29	462	463	465	466	468	469	471	472	474	475
30	477	478	480	481	482	484	485	487	488	490
31	491	492	494	495	496	498	499	501	502	503
32	505	506	507	509	510	511	513	514	515	517
33	518	519	521	522	523	525	526	527	528	530
34	531	532	534	535	536	537	539	540	541	542
35	544	545	546	547	549	550	551	552	553	555
36	556	557	558	559	561	562	563	564	565	567
37	568	569	570	571	572	574	575	576	577	578
38	579	580	582	583	584	585	586	587	588	589
39	591	592	593	594	595	596	597	598	599	601
40	502	603	604	605	606	607	608	609	610	611
41	612	613	614	616	617	618	619	620	621	622
42	623	624	625	626	627	628	629	630	631	632
43	633	634	635	636	637	638	639	640	641	642
44	643	644	645	646	647	648	649	650	651	652
45	653	654	655	656	657	658	659	659	660	661
46	662	663	664	665	666	667	668	669	670	671
47	672	673	673	674	675	676	677	678	679	680
48	681	682	683	683	684	685	686	687	688	689
49	690	691	692	692	693	694	695	696	697	698
50	699	699	700	701	702	703	704	705	705	706
51	707	708	709	710	711	712	713	713	715	715
52	716	716	717	718	719	720	721	722	722	723
53	724	725	725	726	727	728	729	730	730	731
54	732	733	734	734	735	736	737	738	738	739
N	0	1	2	3	4	5	6	7	8	9

THREE-PLACE LOGARITHMS

N	0	1	2	3	4	5	6	7	8	9
55	740	741	741	742	743	744	745	746	747	747
56	748	749	749	750	751	752	752	753	754	755
57	755	756	757	758	758	759	760	761	761	762
58	763	764	764	765	766	767	767	768	769	770
59	770	771	772	773	773	774	775	776	776	777
60	778	778	779	780	781	781	782	783	783	784
61	785	786	786	787	788	788	789	790	791	791
62	792	793	793	794	795	795	796	797	798	798
63	799	800	800	801	802	802	803	804	804	805
64	806	806	807	808	809	810	810	811	811	812
65	813	813	814	814	815	816	816	817	818	818
66	819	820	820	821	822	822	823	824	824	825
67	826	826	827	828	828	829	829	830	831	831
68	832	833	833	834	835	835	836	837	837	838
69	838	839	840	840	841	842	842	843	843	844
70	845	845	846	847	848	848	849	849	850	850
71	851	851	852	853	853	854	854	855	856	856
72	857	857	858	859	859	860	860	861	861	862
73	863	863	864	865	865	866	866	867	868	868
74	869	869	870	871	871	872	872	873	873	874
75	875	875	876	876	877	877	878	879	879	880
76	880	881	882	882	883	883	884	884	885	885
77	886	887	887	888	888	889	889	890	891	891
78	892	892	893	893	894	894	895	896	896	897
79	897	898	898	899	899	900	900	901	902	902
80	903	903	904	904	905	905	906	906	907	907
81	908	909	909	910	910	911	911	912	912	913
82	913	914	914	915	915	916	917	917	918	918
83	919	919	920	920	921	921	922	922	923	923
84	924	924	925	925	926	926	927	927	928	928
85	929	929	930	930	931	932	932	933	933	934
86	934	935	935	936	936	937	937	938	938	939
87	939	940	940	941	941	942	942	943	943	944
88	944	945	945	946	946	946	947	947	948	948
89	949	949	950	950	951	951	952	952	953	953
90	954	954	955	955	956	956	957	957	958	958
91	959	959	960	960	960	961	961	962	962	963
92	963	964	964	965	965	966	966	967	967	968
93	968	968	969	969	970	970	971	971	972	972
94	973	973	974	974	975	975	975	976	976	977
95	977	978	978	979	979	980	980	980	981	981
96	982	982	983	983	984	984	985	985	985	986
97	986	987	987	988	988	989	989	989	990	990
98	991	991	992	992	993	993	993	994	994	995
99	995	996	996	996	997	997	998	998	999	999
00	000	004	008	012	017	021	025	029	033	037
N	0	1	2	3	4	5	6	7	8	9

DB	POWER RATIO
0	1.00
1	1.26
2	1.58
3	2.00
4	2.51
5	3.16
6	3.98
7	5.01
8	6.31
9	7.94
10	10.00
20	100
30	1,000
40	10,000
50	100,000
60	1,000,000
70	10,000,000
80	100,000,000

Power to Decibels. The power output (watts) of any amplifier may be converted into decibels by the following formula:

$$N_{db} = 10 \text{ Log}_{10} \frac{P_1}{P_2} \qquad (2)$$

where N_{db} is the desired power level in decibels; P_1, the output of the amplifier, and P_2, the reference level of 6 milliwatts. The subnumeral, 10, affixed to the logarithm indicates that the log is to be extracted from a log table using 10 as the base, such as the one given here.

Substitute values for the letters in the above formula as in the following:

An amplifier using a 2A5 tube should be able to deliver an undistorted output of 3 watts. How much is this in decibels?

Solution by formula (2)

$$\frac{P_1}{P_2} = \frac{3}{.006} = 500$$

$$10 \times \text{Log } 500 = 10 \times 2.69$$

therefore $10 \times 2.69 = 26.9$ decibels.

Substituting other values for those shown allows any output power to be converted into decibels (in reference to the standard of 6 mw. as 0 level) *provided* that the decibel equivalent is *above* the zero reference level or the power is *not less* than 6 milliwatts.

To solve almost all problems to which the solution will be given in minus decibels, an understanding of *algebraic addition is required.* To add algebraically, it is necessary to observe the plus and minus signs of expressions. (Do not confuse these signs with decibels.) In the succeeding illustrations, notice that the result is obtained sometimes by addition and at other times by subtraction.

(a)	(b)	(c)	(d)
+2	−4	+4	+4
−4	−2	−2	+2
−2	−6	+2	+6

The terms used in (c) are those that apply to decibel calculations.

When the solution to a problem involving logarithms will be in minus decibels (when the power level under consideration is less than 6 milliwatts), note particularly that the characteristic of this logarithm will be prefixed by a minus sign (−). Note also that this sign affects *only* the characteristic; the mantissa remains positive. The mantissa *always* remains positive, regardless of whether the solution of the problem results in a positive or a negative characteristic.

A prefix −1 to a logarithm means that the first significant figure of the number which it represents will be the *first* place to the *right* of the decimal point; −2 means that it will occupy the second place to the right while the first will be filled by a cipher; −3, the third place with two ciphers filling the first and second, and so on.

To multiply a logarithm with a *minus* characteristic and a positive mantissa by another number, each part must be considered separately, multiplied by the number (10 or 20 for decibel calculations), and then the products added algebraically. Thus, in the following illustration:

A preamplifier for a microphone is feeding 1.5 milliwatts into the line going to the regular speech amplifier. What is this power level expressed in decibels? Solution by formula (2):

$$\frac{P_1}{P_2} = \frac{.0015}{.006} = .25$$

Log .25 = −1.397 (from table). Therefore, $10 \times -1.397 = (10 \times -1 = -10) + (10 \times .397 = 3.97)$; adding the products *algebraically* gives −6.03 db.

By substituting other values for those in the above example, any output power *below* 6 milliwatts (zero reference level) can be converted into decibels.

Determining Db Gain or Loss. In using amplifiers, it is a prime requisite to be able to indicate gain or loss in *decibels.* To determine the gain or loss in db employ the following formula:

$$\text{(gain) } N_{db} = 10 \text{ Log } \frac{P_0}{P_1} \qquad (3)$$

$$\text{(loss) } N_{db} = 10 \text{ Log } \frac{P_1}{P_0} \qquad (4)$$

where N_{db} is the number of db gained or lost; P_1, the input power, and P_0, the output power.

Applying, for example, formula (3): Suppose that an intermediate amplifier is being driven by an input power of 0.2 watt and after amplification, the output is found to be 6 watts.

$$\frac{P_0}{P_1} = \frac{6}{.2} = 30$$

$$\text{Log } 30 = 1.48$$

Therefore $10 \times 1.48 = 14.8$ db power gain.

Amplifier Ratings. The technical specifications or rating on power amplifiers should contain the following information: the overall gain in decibels, the power output in watts, the value of the input and output impedances, the input signal level in db, the input signal voltage, and the power output level in decibels.

If the specifications on an amplifier include only the input and output signal levels in db, it then is necessary to calculate how much these values represent in power. The methods employed to determine power levels are not similar to those used in previous calculations. Caution should, therefore, be taken in reading the following explanations, with particular care and attention being paid to the minor arithmetical operations.

The Antilogarithm. To determine a power level from some given decibel value, it is necessary to invert the logarithmic process formerly employed in converting power to decibels. Here, instead of looking for the log of a number, it is now necessary to find the *antilogarithm,* or number corresponding to a given logarithm.

In deriving a number corresponding to a logarithm, it is important that these simple rules be committed to memory: (1) that the figures that form the original number from a corresponding logarithm depend entirely upon the mantissa or decimal part of the log, (2) that the characteristic serves only to indicate where to place the decimal point of the original number, (3) that, if the original number was a whole number, the decimal point would be placed to the extreme right.

The procedure of finding the number corresponding to a logarithm is explained by the following: suppose the logarithm to be 3.574. First, search in the table under any column from 0 to 9 for the numbers of the mantissa 574. If the exact number cannot be found, look for the next *lowest* figure which is nearest to, but less than, the given mantissa. After the mantissa has been located, simply glance immediately to the left to the N column and there will be read the number, 37. This number comprises the first two figures of the number corresponding to the antilog. The third figure of the number will appear at the head of the column in which the mantissa was found. In this instance, the number heading the column will be 5. If the figures have been arranged as they have been found, the number will now be 375.

Now, since the characteristic is 3, there must be four figures to the *left* of the decimal point; therefore, by annexing a cipher, the number becomes 3750; this is the number that corresponds to the logarithm 3.574. If the characteristic were 2 instead of 3, the number would be 375. If the logarithm were -3.574 or -1.574, the antilogs or corresponding numbers would be .00375 and .375 respectively. After a little experience, a person can obtain the number corresponding to a logarithm in a very few seconds.

Converting Decibels to Power. It is always convenient to be able to convert a decibel value to a power equivalent. The formula used for converting decibels into watts is similar in many respects to equation (2), the only difference being that the factor P_1 corresponding to the power level is not known. Usually, the formula for converting decibels into power is written as:

$$N_{db} = 10 \text{ Log } \frac{P_1}{.006} \qquad (5)$$

It is difficult to derive the solution to the above equation because of the expression being written in the reverse. However, by rearranging the various factors, the expression can be simplified to permit easy visualization:

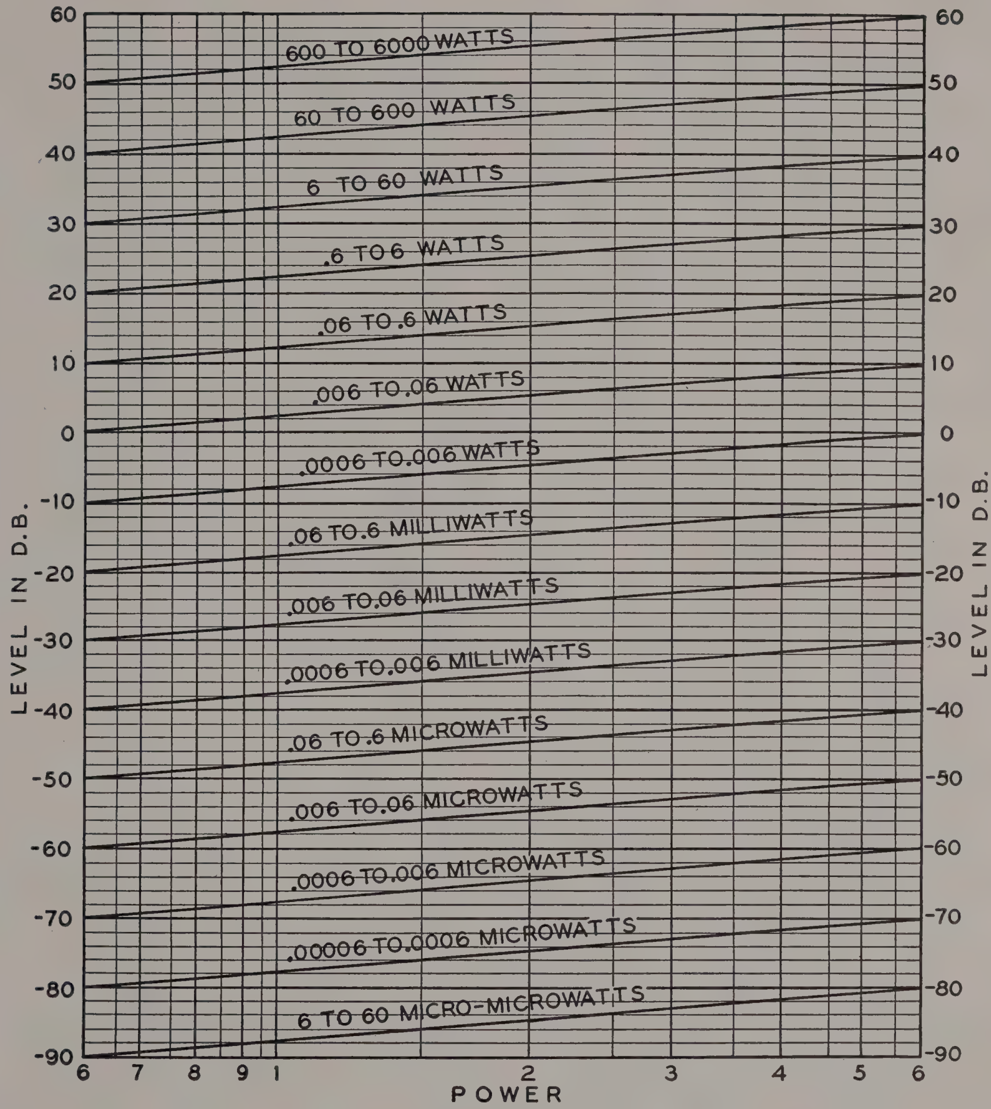
$$P = .006 \times \text{antilog } \frac{N_{db}}{10} \qquad (6)$$

where P is the desired power level; .006, the reference level in milliwatts; N_{db}, the decibels to be converted, and 10, the divisor.

To determine the power level, P, from a decibel equivalent, simply divide the decibel value by 10, then take the number comprising the antilog and multiply it by .006; the product gives the power level of the decibel value.

NOTE: In all problems dealing with the conversion of *minus* decibels to power, it often happens that the decibel value $-N_{db}$, is not always equally divisible by 10. When this is the case, the numerator in the factor $-N_{db}/10$ must be made evenly divisible by the denominator in order to derive the proper power ratio. Note that the value $-N_{db}$ is negative; hence, when dividing by 10, the negative signs must be observed, and the quotient labeled accordingly.

CONVERSION CHART: POWER TO DECIBELS



Based on .006 watts at zero level.

Power levels between 6 micromicrowatts and 6000 watts may be referred to corresponding decibel levels between -90 and 60 db, and vice versa, by means of the above chart. Fifteen ranges are provided. Each curve begins at the same point where the preceding one ends, enabling uninterrupted coverage of the wide db and power ranges with condensed chart. For example: the lowermost curve ends at -80 db or 60 micromicrowatts and the next range starts at the same level. Zero db is taken as 6 milliwatts (.006 watt).

To make the numerator in the value $-N_{db}$ equally divisible by 10, proceed as follows: assume $-N_{db}$ to be the value -38; hence, to make this figure equally divisible by 10, we must add a -2 to it, and, since we have added a negative 2 to it, we must also add a positive 2 to make the net result the same.

Our decibel value now stands, $-40+2$. Dividing both of these figures by 10 (as in equation 6), we have -4 and a plus 0.2. Putting the two of them together, we have -4.2 as our resulting logarithm, with the negative characteristic and positive mantissa required to indicate a number smaller than 1.

While the above discussion applies strictly to negative values, the following examples will clearly show the technique to be followed for almost all practical problems.

(a) The output level of a popular velocity ribbon microphone is rated at -74 db. What is the equivalent in milliwatts?

Solution by equation (6)

$$\frac{-N_{db}}{10} = \frac{-74}{10} \text{ (not equally divisible by 10)}$$

Routine:

$$\begin{array}{rr} -74 & \\ -\ \ 6 & +6 \\ \hline -80 & +6 \end{array}$$

$$\frac{-N_{db}}{10} = \frac{-80 \quad +6}{10} = -8.6$$

Antilog $-8.6 = .00000004$

$.006 \times .00000004 = .00000000024$ watt or 240 micro-microwatts.

(b) This example differs somewhat from that of the foregoing one in that the mantissas are added differently. A low-powered amplifier has an input signal level of -17.3 db. How many milliwatts does this value represent?

Solution by equation (6)

Routine:

$$\begin{array}{rr} -17.3 & \\ -\ \ 2.7 & +2.7 \\ \hline -20.0 & +2.7 \end{array}$$

$$\frac{-N_{db}}{10} = \frac{-20 \quad +2.7}{10} = -2.27$$

Antilog $-2.27 = .0186$

$.006 \times .0186 = .0001116$ watt or .1116 milliwatts.

Voltage Amplifiers. When plans are being drafted contemplating the design of power amplifiers, it is essential that the following data be determined: first, the input and output signal levels to be used; second, the size of the power tubes that will adequately deliver sufficient undistorted output; third, the input signal voltage that must be applied to the amplifier to deliver the desired output. This last requirement is the most important in the design of voltage amplifiers.

The voltage set-up in a transformer-coupled amplifier depends chiefly upon the μ of the tubes and the turns ratio of the interstage coupling transformers. The step-up value in any amplifier is calculated bv multiplying the step-up factor of each voltage amplifying or step-up device. Thus, for example, if an amplifier were designed having an output transformer with a ratio of 3:1 coupled to a tube having a μ of 7, the voltage step-up would be approximately 3 times 7, or 21. It is seldom that the total product will be exactly the figure derived, because it is not possible to realize amplification equal to the full μ of the tube.

Decibel-Voltage Ratios. From the voltage gain in an amplifier, it is possible to calculate the input and output signal levels and at the same time be able to determine at what level the input signal must be in order to obtain the desired output. By converting voltage ratios into decibels, power levels can be determined. Hence, to find the gain in db when the input and output voltages are known, the following expression is used:

$$\text{(gain) } N_{db} = 20 \text{ Log } \frac{E_1}{E_2} \quad (7)$$

where E_1, is the output voltage, and E_2, the input voltage.

Employing the above equation in a practical problem, note the logarithm is multiplied by 20 instead of by 10, as in previous examples. For instance:

A certain one-stage amplifier consists of the following parts: 1 input transformer, ratio 2:1, and 1 output tube having a μ of 95. Determine the gain in decibels, assuming an input voltage of 1 volt.

Solution by equation (7)

$$2 \times 95 = 190 \text{ voltage gain}$$

$$\text{therefore, } \frac{E_1}{E_2} = \frac{190}{1} = 190$$

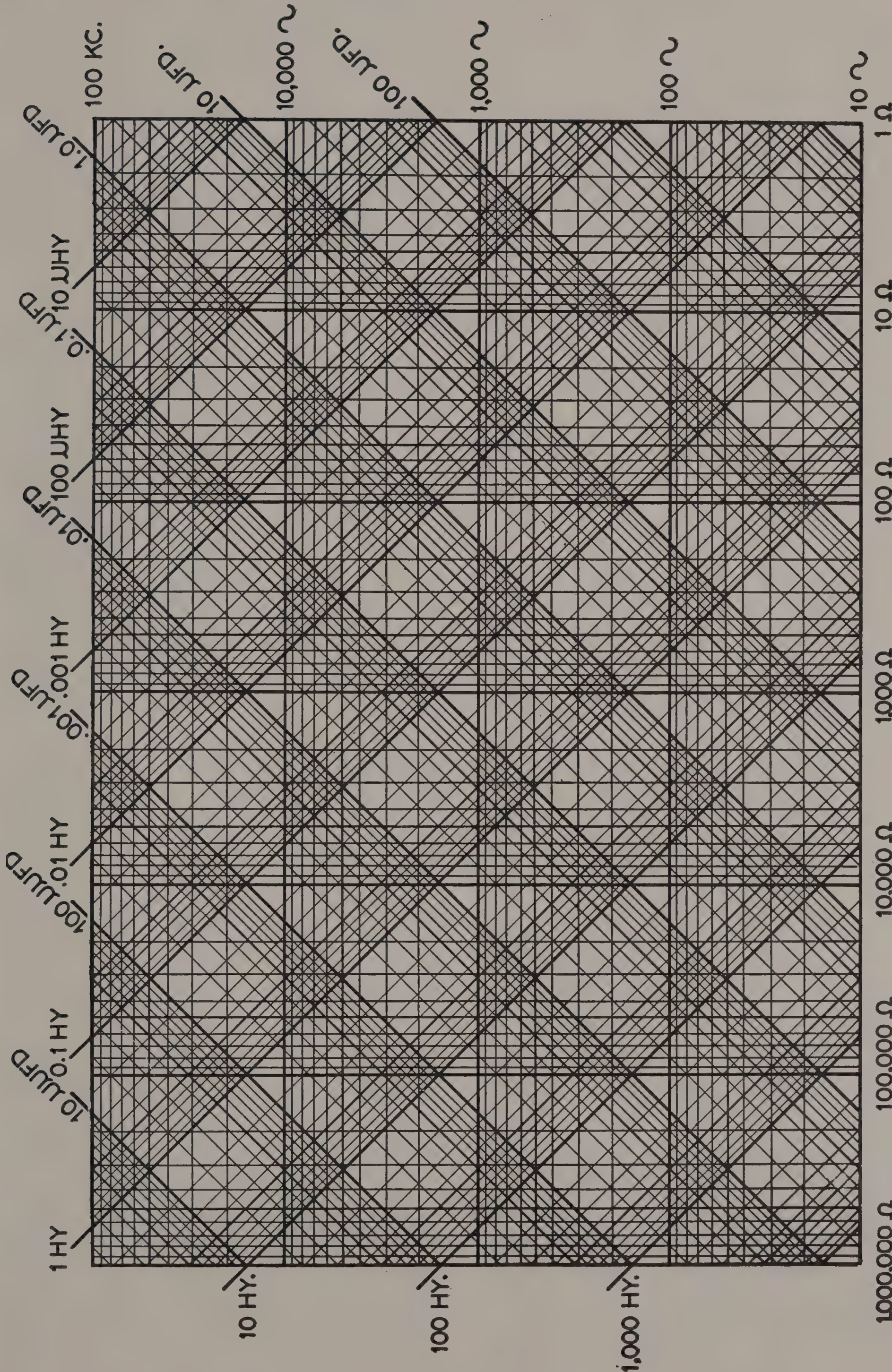
$$\text{Log } 190 = 2.278$$

$20 \times 2.278 = 45.56$ decibels gain.

To reverse the above and convert decibels to voltage ratios, use the following expression:

REACTANCE-FREQUENCY CHART
FOR AUDIO FREQUENCIES

See text for applications and instructions for use.



$$E \text{ (gain)} = \text{antilog } \frac{N_{db}}{20} \quad (8)$$

where E is the voltage gain (power ratio); N_{db}, the decibels, and 20, the divisor.

To find the gain, simply divide the decibels by 20, then extract the antilog from the quotient; the result gives the voltage ratio.

Input Voltages. In designing power amplifiers, it is paramount to have *exact* knowledge of the magnitude of the input signal voltage necessary to drive the output power tubes to maximum undistorted output.

To determine the required input voltage, take the *peak voltage* necessary to drive the grid of the last class A amplifier tube to maximum output, and divide this figure by the total overall gain *preceding this stage.*

Computing Specifications. From the preceding explanations, the following data can be computed with a very high degree of accuracy:

(1) Voltage amplification
(2) Overall gain in db
(3) Output signal level in db
(4) Input signal level in db
(5) Input signal level in watts
(6) Input signal voltage.

Push-Pull Amplifiers. To double the output of any cascade amplifier, it is only necessary to connect in push-pull the last amplifying stage, and replace the interstage and output transformers with push-pull types.

To determine the voltage gain (voltage ratio) of a push-pull amplifier, take the ratio of one *half* of the secondary winding of the push-pull transformer and multiply it by the μ of one of the output tubes in the push-pull stage; the product, *when doubled,* will be the voltage amplification, or step-up.

Acoustically (that is, from the loudspeaker standpoint) it takes approximately 3 db before any change in the volume of sound is noted. This is because the intensity of sound as heard by the ear varies logarithmically with the acoustic power. For practical purposes, it is only necessary to remember that if two sounds differ in physical intensity by less than 3 db, they sound practically alike.

Preamplifiers. Preamplifiers are employed to raise low input signal levels up to some required input level of another intermediate or succeeding amplifier. For example: if an amplifier was designed to operate at an input level of -30 db, and instead a considerably lower input level were used, a preamplifier would then have to be designed to bring the low input signal up to the rated input-signal level of -30 db to obtain the full undistorted output from the power tubes in the main amplifier. The amount of gain necessary to raise a low input-signal level up to another level may be determined by the following equation:

$$E \text{ (gain)} = \text{antilog } \frac{N_{db1} - N_{db2}}{20} \quad (9)$$

where E is the voltage step-up or gain; N_{db1}, the input signal level of the preamplifier or the new input signal level; N_{db2}, the input signal level to the intermediate amplifier; and 20, the divisor.

Reactance Calculations

In audio frequency calculations, an accuracy to better than a few per cent is seldom required, and when dealing with calculations involving inductance, capacitance, resonant frequency, etc., it is much simpler to make use of a reactance-frequency chart such as that on page 552 rather than to wrestle with a combination of unwieldy formulas. From this chart it is possible to determine the reactance of a condenser or coil if the capacitance or inductance is known, and vice versa. It follows from this that resonance calculations can be made directly from the chart, because resonance simply means that the inductive and capacitive reactances are equal. The capacity required to resonate with a given inductance, or the inductance required to resonate with a given capacity, can be taken directly from the chart.

While the chart may look somewhat formidable to one not familiar with charts of this type, its application is really quite simple, and can be learned in a short while. The following example should clarify its interpretation.

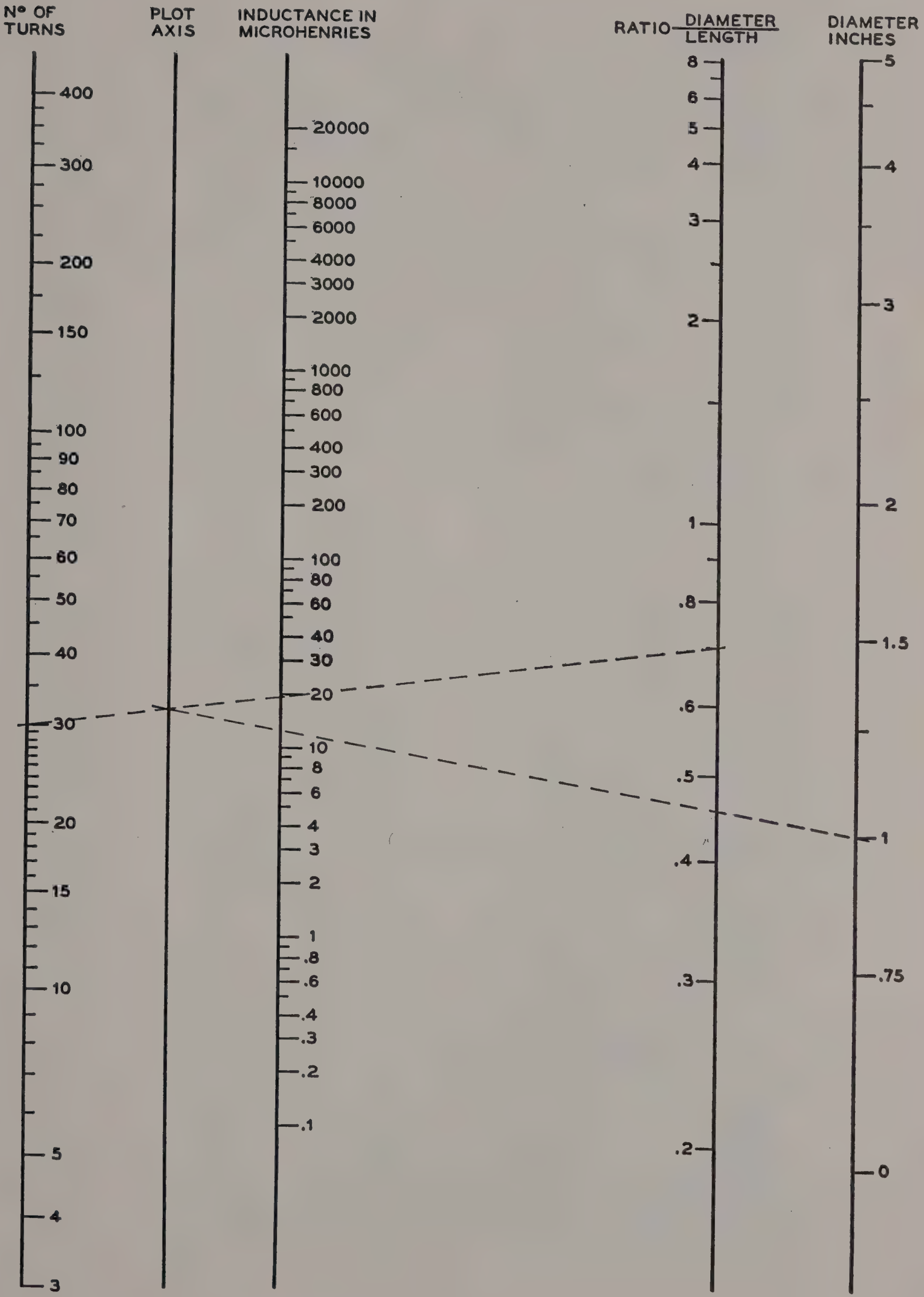
For instance, following the lines to their intersection, we see that 0.1 hy. and 0.1 μfd. intersect at approximately 1,500 cycles and 1000 ohms. Thus, the reactance of either the coil or condenser taken alone is about 1000 ohms, and the resonant frequency about 1,500 cycles.

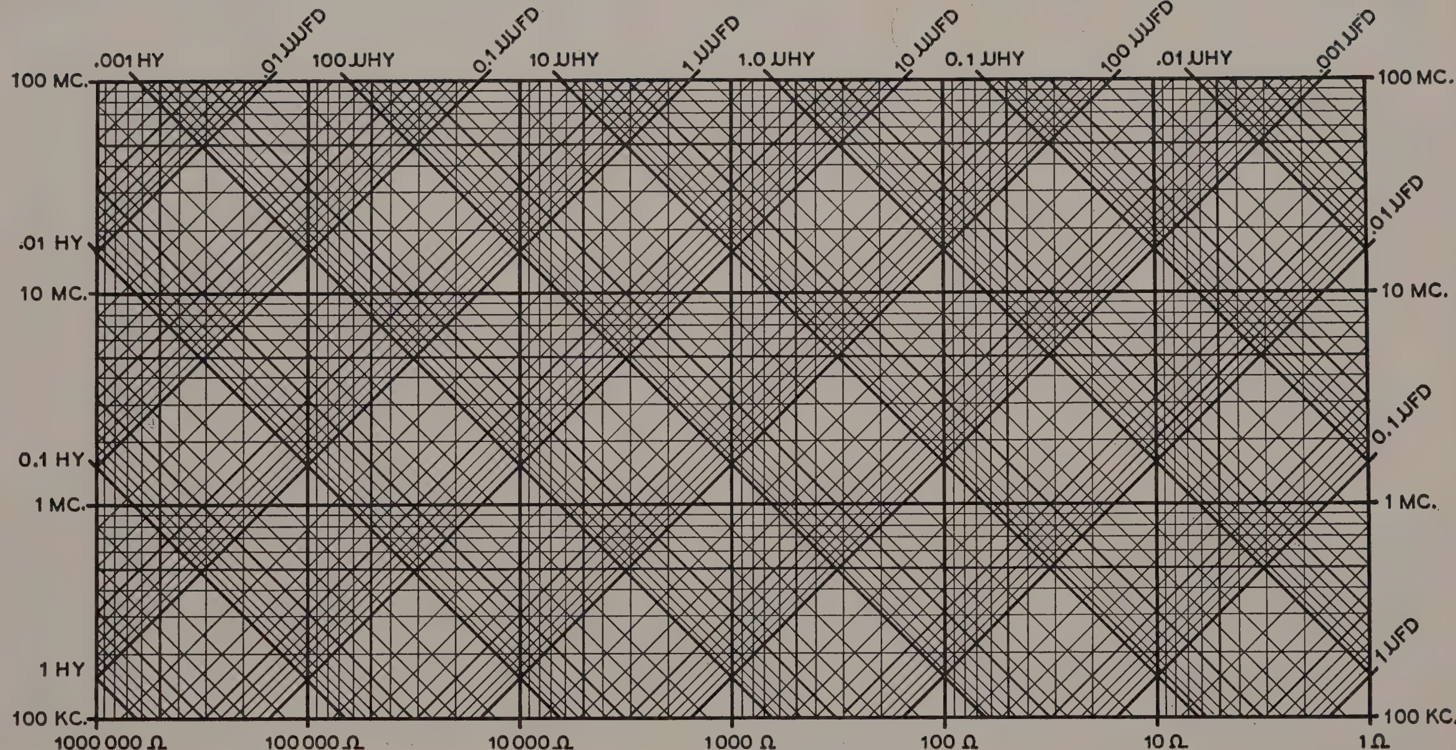
To find the reactance of 0.1 hy. at say 10,000 cycles, simply follow the inductance line diagonally up towards the upper left till it intersects the horizontal 10,000 kc. line. Following vertically downward from the point of intersection, we see that the reactance at this frequency is about 6000 ohms.

To facilitate use of the chart and to avoid errors, simply keep the following in mind: The vertical lines indicate reactance in ohms, the horizontal lines always indicate the frequency, the diagonal lines sloping to the lower right represent inductance, and the diagonal lines sloping toward the lower left indicate

COIL CALCULATOR NOMOGRAPH

For single layer solenoid coils, any wire size. See text for instructions.





REACTANCE-FREQUENCY CHART FOR R. F.

This chart is used in conjunction with the nomograph on opposite page for radio frequency tank coil computations.

capacitance. Also remember that the scale is *logarithmic.* For instance, the next horizontal line above 1000 cycles is *2000* cycles. Note that there are 9, not 10, divisions between the heavy lines. This also should be kept in mind when interpolating between lines when best possible accuracy is desired; halfway between the line representing 200 cycles and the line representing 300 cycles is *not* 250 cycles, but approximately 230 cycles. The 250 cycle point is approximately 0.7 of the way between the 200 cycle line and the 300 cycle line, rather than halfway between.

Use of the chart need not be limited by the physical boundaries of the chart. For instance, the 10-$\mu\mu$fd. line can be extended to find where it intersects the 100-hy. line, the resonant frequency being determined by projecting the intersection horizontally back on to the chart. To determine the reactance, the logarithmic ohms scale must be extended.

R. F. Tank Circuit Calculations. When winding coils for use in radio receivers and transmitters, it is desirable to be able to determine in advance the full coil specifications for a given frequency. Likewise, it often is desired to determine how much capacity is required to resonate a given coil so that a suitable condenser can be used.

Fortunately, extreme accuracy is not required, except where fixed capacitors are used across the tank coil with no provision for trimming the tank to resonance. Thus, even though it may be necessary to estimate the stray circuit capacity present in shunt with the tank capacity, and to take for granted the likelihood of a small error when using a chart instead of the formula upon which the chart was based, the results will be sufficiently accurate in most cases, and in any case give a reasonably close point from which to start "pruning."

The inductance required to resonate with a certain capacitance is given in the chart on page 550. The data previously given on using the audio frequency reactance chart also apply to the r.f. chart. By means of the r.f. chart, the inductance of the coil can be determined, or the capacitance determined if the inductance is known. When making calculations, be sure to allow for stray circuit capacity, such as tube interelectrode capacity, wiring, sockets, etc. This will normally run from 5 to 25 $\mu\mu$fd., depending upon the components and circuit.

To convert the inductance in microhenries to physical dimensions of the coil, or vice versa, the nomograph chart on page 549 is used. A pin and a straightedge are required. The inductance of a coil is found as follows:

The straightedge is placed from the correct point on the turns column to the correct point on the diameter-to-length ratio column, the latter simply being the diameter divided by the length. Place the pin at the point on the plot axis column where the straightedge crosses it. From this point lay the straightedge to the correct point on the diameter column. The point where the straightedge intersects the inductance column will give the inductance of the coil.

From the chart, we see that a 30 turn coil having a diameter-to-length ratio of 0.7 and a diameter of 1 inch has an inductance of approximately 12 microhenries. Likewise any one of the four factors may be determined if the other three are known. For instance, to determine the number of turns when the desired inductance, the D/L ratio, and the diameter are known, simply work backwards from the example given. In all cases, remember that the straightedge reads either turns and D/L ratio, *or* it reads inductance and diameter. It can read no other combination.

The actual wire size has negligible effect upon the calculations for commonly used wire sizes (no. 10 to no. 30). The number of turns of insulated wire that can be wound per inch (solid) will be found in the copper wire table on page 334.

Radio Laws

Pertinent Extracts from the Communications Act of 1934, as Amended; International Radiotelegraph Conference, Madrid, 1932; International Radio Regulations (Cairo Revision, 1938).

Extracts of the Communications Act of 1934, as amended.

Section 1. For the purpose of regulating interstate and foreign commerce in communication by wire and radio so as to make available, so far as possible, to all the people of the United States a rapid, efficient, Nation-wide, and world-wide wire and radio communication service with adequate facilities at reasonable charges, for the purpose of the national defense, for the purpose of promoting safety of life and property through the use of wire and radio communication, and for the purpose of securing a more effective execution of this policy by centralizing authority heretofore granted by law to several agencies and by granting additional authority with respect to interstate and foreign commerce in wire and radio communication, there is hereby created a Commission to be known as the "Federal Communications Commission," which shall be constituted as hereinafter provided and which shall execute and enforce the provisions of this act.

Sec. 301. It is the purpose of this Act, among other things, to maintain the control of the United States over all the channels of interstate and foreign radio transmission; and to provide for the use of such channels, but not the ownership thereof, by persons for limited periods of time, under licenses granted by Federal authority, and no such license shall be construed to create any right, beyond the terms, conditions, and periods of the license. No person shall use or operate any apparatus for the transmission of energy or communications or signals by radio (a) from one place in any Territory or possession of the United States or in the District of Columbia to another place in the same Territory, possession, or district; or (b) from any State, Territory, or possession of the United States, or from the District of Columbia to any other State, Territory, or possession of the United States; or (c) from any place in any State, Territory, or possession of the United States, or in the District of Columbia, to any place in any foreign country or to any vessel; or (d) within

AMATEUR REGULATIONS

Because of the present unsettled status of amateur regulations, no regulations pertaining specifically to amateurs are included. There would be no way for the reader to know whether they had been revised or supplemented since this book went to press. Hence, for rules and regulations pertaining to amateur stations and operators, the reader should write to the Government Printing Office, Washington, D. C.

any State when the effects of such use extend beyond the borders of said State, or when interference is caused by such use or operation with the transmission of such energy, communications, or signals from within said State to any place beyond its borders, or from any place beyond its borders to any place within said State, or with the transmission or reception of such energy, communications, or signals from and/or to places beyond the borders of said State; or (e) upon any vessel or aircraft of the United States; or (f) upon any other mobile stations within the jurisdiction of the United States, except under and in accordance with this Act and with a license in that behalf granted under the provisions of this Act.

Sec. 303. Except as otherwise provided in this Act, the Commission from time to time, as public convenience, interest, or necessity requires, shall—

(1) Have authority to prescribe the qualifications of station operators, to classify them according to the duties to be performed, to fix the forms of such licenses, and to issue them to such citizens of the United States as the Commission finds qualified;

(m) (1) Have authority to suspend the license of any operator upon proof sufficient to satisfy the Commission that the licensee—

(A) Has violated any provision of any Act, treaty, or convention binding on the United States which the Commission is authorized to administer, or any regulation made by the Commission under any such Act, treaty, or convention; or

(B) Has failed to carry out a lawful order of the master or person lawfully in charge of the ship or aircraft on which he is employed; or

(C) Has willfully damaged or permitted radio apparatus or installations to be damaged; or

(D) Has transmitted superfluous radio communications or signals or communications containing profane or obscene words, language, or meaning, or has knowingly transmitted—

(1) False or deceptive signals or communications, or

(2) A call signal or letter which has not been assigned by proper authority to the station he is operating; or

(E) Has willfully or maliciously interfered with any other radio communications or signals; or

(F) Has obtained or attempted to obtain, or has assisted another to obtain or attempt to obtain, an operator's license by fraudulent means.

(2) No order of suspension of any operator's license shall take effect until fifteen days' notice in writing thereof, stating the cause for the proposed suspension, has been given to the operator licensee who may make written application to the Commission at any time within said fifteen days for a hearing upon such order. The notice to the operator licensee shall not be effective until actually received by him, and from that time he shall have fifteen days in which to mail the said application. In the event that physical conditions prevent mailing of the application at the expiration of the fifteen-day period, the application shall then be mailed as soon as possible thereafter, accompanied by a satisfactory explanation of the delay. Upon receipt by the Commission of such application for hearing, said order of suspension shall be held in abeyance until the conclusion of the hearing which shall be conducted under such rules as the Commission may prescribe. Upon the conclusion of said hearing the Commission may affirm, modify, or revoke said order of suspension.

(n) Have authority to inspect all radio installations associated with stations required to be licensed by any Act or which are subject to the provisions of any Act, treaty, or convention binding on the United States, to ascertain whether in construction, installation, and operation they conform to the requirements of the rules and regulations of the Commission, the provisions of any Act, the terms of any treaty or convention binding on the United States, and the conditions of the license or other instrument of authorization under which they are constructed, installed, or operated.

(r) Make such rules and regulations and prescribe such restrictions and conditions, not inconsistent with law, as may be necessary to carry out the provisions of this Act, or any international radio or wire communications treaty or convention, or regulations annexed thereto, including any treaty or convention insofar as it relates to the use of radio, to which the United States is or may hereafter become a party.

Sec. 318. The actual operation of all transmitting apparatus in any radio station for which a station license is required by this Act shall be carried on only by a person holding an operator's license issued hereunder, and no person shall operate any such apparatus in such station except under and in accordance with an operator's license issued to him by the Commission: *Provided, however,* That the Commission if it shall find that the public interest, convenience, or necessity will be served thereby may waive or modify the foregoing provisions of this section for the operation of any station except (1) stations for which licensed operators are required by international agreement, (2) stations for which licensed operators are required for safety purposes, (3) stations en-

gaged in broadcasting and (4) stations operated as common carriers on frequencies below thirty thousand kilocycles: *Provided further,* That the Commission shall have power to make special regulations governing the granting of licenses for the use of automatic radio devices and for the operation of such devices.

Sec. 321. (a) The transmitting set in a radio station on shipboard may be adjusted in such a manner as to produce a maximum of radiation, irrespective of the amount of interference which may thus be caused, when such station is sending radio communications or signals of distress and radio communication relating thereto.

(b) All radio stations, including Government stations and stations on board foreign vessels when within the territorial waters of the United States, shall give absolute priority to radio communications or signals relating to ships in distress; shall cease all sending on frequencies which will interfere with hearing a radio communication or signal of distress, and, except when engaged in answering or aiding the ship in distress, shall refrain from sending any radio communications or signals until there is assurance that no interference will be caused with the radio communications or signals relating thereto, and shall assist the vessel in distress, so far as possible, by complying with its instructions.

Sec. 322. Every land station open to general public service between the coast and vessels or aircraft at sea shall, within the scope of its normal operations, be bound to exchange radio communications or signals with any ship or aircraft station at sea; and each station on shipboard or aircraft at sea shall, within the scope of its normal operations, be bound to exchange radio communications or signals with any other station on shipboard or aircraft at sea or with any land station open to general public service between the coast and vessels or aircraft at sea: *Provided,* That such exchange of radio communication shall be without distinction as to radio systems or instruments adopted by each station.

Sec. 325. (a) No person within the jurisdiction of the United States shall knowingly utter or transmit, or cause to be uttered or transmitted, any false or fraudulent signal of distress, or communication relating thereto, nor shall any broadcasting station rebroadcast the program or any part thereof of another broadcasting station without the express authority of the originating station.

Sec. 326. Nothing in this Act shall be understood or construed to give the Commission the power of censorship over the radio communications or signals transmitted by any radio station, and no regulation or condition shall be promulgated or fixed by the Commission which shall interfere with the right of free speech by means of radio communication. No person within the jurisdiction of the United States shall utter any obscene, indecent, or profane language by means of radio communication.

Sec. 358. The radio installation, the operators, the regulation of their watches, the transmission and receipt of messages, and the radio service of the ship except as they may be regulated by law or international agreement, or by rules and regulations made in pursuance thereof, shall in the case of a ship of the United States be under the supreme control of the master.

Sec. 501. Any person who willfully and knowingly does or causes or suffers to be done any act, matter, or thing, in this Act prohibited or declared to be unlawful, or who willfully and knowingly omits or fails to do any act, matter, or thing in this Act required to be done, or willfully and knowingly causes or suffers such omission or failure, shall, upon conviction thereof, be punished for such offense, for which no penalty (other than a forfeiture) is provided herein, by a fine of not more than $10,000 or by imprisonment for a term of not more than two years, or both.

Sec. 502. Any person who willfully and knowingly violates any rule, regulation, restriction, or condition made or imposed by the Commission under authority of this Act, or any rule, regulation, restriction, or condition made or imposed by any international radio or wire communications treaty or convention, or regulations annexed thereto, to which the United States is or may hereafter become a party, shall, in addition to any other penalties provided by law, be punished, upon conviction thereof, by a fine of not more than $500 for each and every day during which such offense occurs.

Sec. 605. No person receiving or assisting in receiving, or transmitting, or assisting in transmitting, any interstate or foreign communication by wire or radio shall divulge or publish the existence, contents, substance, purport, effect, or meaning thereof, except through authorized channels of transmission or reception, to any person other than the addressee, his agent, or attorney, or to a person employed or authorized to forward such communication to its destination, or to proper accounting or distributing officers of the various communicating centers over which the communication may be passed, or to the master of a ship under whom he is serving, or in response to a subpoena issued by a court of competent jurisdiction, or on demand of other lawful authority; and no person not being authorized by the

sender shall intercept any communication and divulge or publish the existence, contents, substance, purport, effect, or meaning of such intercepted communication to any person; and no person not being entitled thereto shall receive or assist in receiving any interstate or foreign communication by wire or radio and use the same or any information therein contained for his own benefit or for the benefit of another not entitled thereto; and no person having received such intercepted communication or having become acquainted with the contents, substance, purport, effect, or meaning of the same or any part thereof, knowing that such information was so obtained, shall divulge or publish the existence, contents, substance, purport, effect, or meaning of the same or any part thereof, or use the same or any information therein contained for his own benefit or for the benefit of another not entitled thereto: *Provided,* That this section shall not apply to the receiving, divulging, publishing, or utilizing

DANGER—HIGH VOLTAGE

The high voltage power supplies even in a low-power transmitter are potentially lethal. They are also potential fire hazards. Pages could be written on "don'ts" and precautionary measures, but the important thing is to *use your head;* don't fool with any part of your transmitter or power supply unless you know exactly what you are doing and have your mind on what you are doing.

Not only should your transmitter installation be so arranged to minimize the danger of accidental shock for your own safety, but also because "haywire" installations that do not pass the underwriters' rules will invalidate your fire insurance. You have no claim against the insurance company if they can prove that the installation did not meet underwriters' specifications.

Some of the most important things to remember in regard to the high voltage danger are the following:

Do not rely upon bleeders to discharge your filter condensers; short the condenser with an insulated-handle screwdriver before handling any of the associated circuits. Bleeders occasionally blow out, and good filter condensers hold a charge a long time.

Beware of "zero adjuster" devices on meters placed in positive high-voltage leads. Also be careful of dial set screws if the rotor shaft of the condenser is "hot." Both of these situations represent poor practice to begin with.

Don't touch any transmitter components without first turning off all switches. If you do insist on making coupling adjustments, etc., with the power on (very bad practice), keep ONE HAND BEHIND YOU.

Do not work on the high-voltage circuits or make adjustments where it is necessary to reach inside the transmitter UNLESS SOMEONE ELSE IS PRESENT. Ninety per cent of the deaths of amateurs due to electrocution could have been prevented if someone had been present to kill the high voltage or remove the victim and to call the doctor and administer first aid before he arrived.

High-voltage gear should be so fixed that small children cannot manipulate the switches or come in contact with any of the wires or components. Either keep the radio room or gear under lock and key or else provide an "interlock" system whereby all primary circuits are broken when the transmitter cabinet is opened.

Familiarize yourself with the latest approved methods of first aid treatment for electrical shock. It may enable you to save a life some time.

Don't attempt to hurry too much if a companion comes in contact with high voltage and cannot extricate himself. Act quickly but do not act without deliberation or you may be in as bad a fix as the person you are trying to help. Do not touch the victim with your bare hands if things are wet. Otherwise, it is safe to grab him by a loose fold of clothing to pull him free, first making sure that you are well-insulated from anything grounded. Turning off the voltage is simpler, when possible. However, do not waste precious moments dashing around trying to discover how to open the circuit. If you do not already know, try to remove the victim if it can be done safely.

A main primary switch at the entrance to the radio room, killing all primary circuits, will reduce the fire hazard and help your peace of mind, provided you make it an iron-clad rule always to throw the switch when leaving the room.

Beware of strange equipment. It may contain unconventional wiring or circuits. Do not take for granted that it is wired the way you would do it.

FOR THOSE WHO WANT THE BEST

HYTRON OFFERS

GREATER VALUE AND GREATER EFFICIENCY IN ITS TRIODES, U. H. F. TRIODES AND R. F. BEAM TETRODES

HY75 $3.95

Filament	*6.3 volts @ 2.5 amps.
Plate dissipation (max.)	15 watts
Plate	450 max. volts & 100 max. ma.

Nominal Output	Modulated	Unmodulated
56 megacycles	24	33 watts
112 megacycles	16	19 watts
224 megacycles	12	15 watts

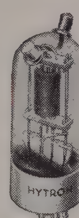
HY615–HY114B

HY114B $2.25 HY615 $2.25

	HY114B	HY615
Filament potential	*1.25 v.	6.3 v.
Filament current	0.145 a.	0.17 a.
Plate potential (max.)	180 v.	300 v.
Plate current (max.)	15 ma.	20 ma.
Plate dissipation (max.)	2 w.	3.5 w.
Class C output	2 w.	4.5 w.

HY30Z $2.75 HY31Z $3.50

	HY30Z	HY31Z
Filament potential	*6.3 v.	*6.3 v.
Filament current	2.25 a.	2.5 a.
Plate potential (max.)	850 v.	500 v.
Plate current (max.)	90 ma.	150 ma.
Plate dissipation (max.)	30 w.	30 w.
Class C output	58 w.	56 w.
Class B audio output	110 w.	51 w.



HY75

HY30Z

HY40 $3.75 HY40Z $3.75

	HY40	HY40Z
Filament potential	7.5 v.	7.5 v.
Filament current	2.25 a.	2.5 a.
Amp. factor	25	80
Plate dissipation (max.)	40 w.	40 w.
Plate	1000 max. v. & 125 max. ma.	
Class C output		94 w.
Class B output (2 tubes)		185 w.

HY51A HY51B HY51Z $4.75

	HY51A, Z	HY51B
Filament potential	7.5 v.	10 v.
Filament current	3.5 a.	2.25 a.
Amp. factor	25–85	25
Plate dissipation (max.)	65 w.	65 w.
Plate	1000 max. v. & 175 max. ma.	
Class C output		131 w.
Class B output (2 tubes)		285 w.

HY31Z

HY51Z

HY24 $1.50 801A/801 $2.50

	HY24	801A
Filament potential	*2.0 v.	7.5 v.
Filament current	0.13 a.	1.25 a.
Plate potential (max.)	180 v.	600 v.
Plate current (max.)	20 ma.	70 ma.
Plate dissipation (max.)	2.0 w.	20 w.
Class C output	2.7 w.	30 w.

All ratings are for continuous-duty service (CCS).

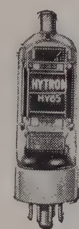
HY63 $2.50 HY67 $7.75

	HY63	HY67
Filament potential	*1.25 or 2.5 v.	*6.3 or 12.6 v.
Filament current	0.22 or 0.11 a.	4 or 2 a.
Plate potential (max.)	200 v.	1250 v.
Plate current (max.)	20 ma.	175 ma.
Plate dissipation (max.)	3 w.	65 w.
Class C output	3.0 w.	152 w.

HY67

HY65 $3.00 HY69 $3.95

	HY65	HY69
Filament potential	*6.3 v.	*6.3 v.
Filament current	0.8 a.	1.6 a.
Plate potential (max.)	450 v.	600 v.
Plate current (max.)	63 ma.	100 ma.
Plate dissipation (max.)	15 w.	40 w.
Class C output	19 w.	42 w.



HY65

HY60 $2.75 HY61/807 $3.50

	HY60	HY61/807
Filament potential	6.3 v.	6.3 v.
Filament current	0.5 a.	0.9 a.
Plate potential (max.)	425 v.	600 v.
Plate current (max.)	60 ma.	100 ma.
Plate dissipation (max.)	15 w.	25 w.
Class C output	16 w.	40 w.

HY866 Jr. $1.05 866A/866 $1.50

	HY866 Jr.	866A/866
Filament potential	2.5 v.	2.5 v.
Filament current	2.5 a.	5.0 a.
Peak inverse potential	5000 v.	10000 v.
Peak plate current	500 ma.	1000 ma.
Max. D.C. output pot.	1575 v.	3165 v.
Max. D.C. Cur. (2 tubes)	250 ma.	500 ma.

HY69

6V6GTX $1.05 6L6GX $1.25

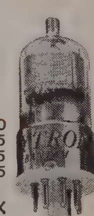
	6V6GTX	6L6GX
Filament potential	6.3 v.	6.3 v.
Filament current	0.5 a.	0.9 a.
Plate potential (max.)	300 v.	500 v.
Plate current (max.)	60 ma.	90 ma.
Plate dissipation (max.)	15 w.	21 w.
Class C output	12 w.	30 w.

866 Jr.

CERAMIC-BASE BANTAMS

6A8GTX	$0.95	6K8GTX	$1.30
6J5GTX	0.95	6SA7GTX	1.05
6J7GTX	0.95	6SJ7GTX	1.05
6K7GTX	0.95	6SK7GTX	1.05

Use specially-selected Hytron ceramic-based GTX tubes for low-loss reception and transmission.



GTX

*Instant-heating filament

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HYTRON RADIO TUBES

A DIVISION OF HYTRON CORP.

Manufacturers of Radio Tubes Since 1921

the contents of any radio communication broadcast, or transmitted by amateurs or others for the use of the general public, or relating to ships in distress.

INTERNATIONAL TELECOMMUNICATION CONVENTION, MADRID, 1932

Article 24

§ 1. The contracting governments agree to take all the measures possible, compatible with the system of telecommunication used, with a view to insuring the secrecy of international correspondence.

* * * *

Article 34

§ 1. Stations carrying on radio communications in the mobile service shall be bound, within the scope of their normal operation, to exchange radio communications with one another irrespective of the radio system they have adopted.

Article 35

§ 1. All stations, regardless of their purpose, must, so far as possible, be established and operated in such a manner as not to interfere with the radio services or communications of either the other contracting governments, or the private operating agencies recognized by these contracting governments and of other duly authorized operating agencies which carry on radio-communication service.

Article 36

Stations participating in the mobile service shall be obliged to accept, with absolute priority, distress calls and messages regardless of their origin, to reply in the same manner to such messages, and immediately to take such action in regard thereto as they may require.

Article 37

The contracting governments agree to take the steps required to prevent the transmission or the putting into circulation of false or deceptive distress signals or distress calls, and the use, by a station, of call signals which have not been regularly assigned to it.

GENERAL RADIO REGULATIONS (CAIRO REVISION, 1938)

ANNEXED TO THE INTERNATIONAL TELECOMMUNICATIONS CONVENTION (MADRID 1932)

Article 2

44 The administrations agree to take the necessary measures to prohibit and prevent:

45 (a) the unauthorized interception of radio communications not intended for the general use of the public;

46 (b) the divulging of the contents or of the mere existence, the publication or any use whatever, without authorization, of the radio communication mentioned in No. 45.

Article 3

47 § 1. (1) No transmitting station may be established or operated by any person or by any enterprise whatever without a special license issued by the government of the country to which the station in question is subject.

* * * *

Article 6

69 § 1. The waves emitted by a station must be kept on the authorized frequency as exactly as the state of the art permits, and their radiation must be as free as practically possible from all emissions not essential to the type of communication carried on.

71 § 2. (1) The state of the art in the various cases of operation is defined in appendixes 1, 2, and 3, concerning the exactitude of the frequency, the level of harmonics, and the width of the frequency band occupied.

* * * *

76 § 3. (1) The administrations shall frequently check the waves emitted by the stations under their jurisdiction to determine whether or not they comply with the provisions of the present Regulations.

* * * *

Article 9

203 § 2. The frequency of emission of mobile stations shall be verified as often as possible by the inspection service to which they are subject.

* * * *

Article 11

276 § 1. The radio service of a mobile station shall be placed under the supreme authority of the master or the person responsible for the ship, aircraft, or any other vehicle carrying the mobile station.

* * * *

278 § 3. The master or responsible person as well as any persons who may have knowledge of the text or simply the existence of radiotelegrams, or of any information acquired by means of the radio service, shall be bound by the obligation to observe and insure the secrecy of the correspondence.

Article 12

279 § 1. (1) The competent governments or administrations of countries where a mobile station calls, may demand the production of

the license. The operator of the mobile station or the person responsible for the station must submit to this verification. The license must be kept in such a way that it may be furnished without delay. However, the production of the license may be replaced by a permanent posting in the station, of a copy of the license certified by the authority which has granted it.

* * * *

Article 17

374 § 2. (1) Before transmitting, any station must keep watch over a sufficient interval to assure itself that it will cause no harmful interference with the transmissions being made within its range; if such interference is likely, the station shall await the first stop in the transmission which it may disturb.

* * * *

Article 22

525 § 1. (1) The transmission of unnecessary or unidentified signals or correspondence shall be forbidden to all stations.

527 (2) Tests and experiments shall be permitted in mobile stations if they do not interfere with the service of other stations. As for stations other than mobile stations, each administration shall judge, before authorizing them, whether or not the proposed tests or experiments are likely to interfere with the service of other stations.

* * * *

EMERGENCY WORK

In many of the larger cities, various individuals and clubs have prepared themselves for any emergency that might arise by constructing and keeping in readiness self-powered communication equipment, for use in case of power failure. Such foresightedness is to be commended.

On the other hand, disaster is bound to strike at times in localities where no such methodical preparation has been made. In such an event, amateurs with a reasonable amount of technical training and experience should be able to establish communication within a couple of hours with no other facilities than an automobile tool kit, two or three auto radio receivers, and a couple of auto batteries. The latter are universally available, and a high percentage of the newer cars are now equipped with auto radios. If the supply of auto batteries is limited, they can be recharged by one of the autos from which they were commandeered.

One receiver can be torn up to provide parts and a power source for the transmitter. The output tube will make a good oscillator, either crystal or "t.n.t."

Another set can be torn into and the coils cut down until with the condenser plates entirely unmeshed they "hit" around 70 meters. Both the 80 and 160 meter bands can then be covered in grand style if a good outside antenna and a ground connection are used.

By wrapping a piece of insulated wire around both plate and grid leads of an i.f. tube (assuming the set is a superheterodyne, as most are), the stage can be made to oscillate, thus providing makeshift yet satisfactory heterodyne reception of c.w. signals. If the a.v.c. action bothers on c.w., either use a very short antenna or else ground the a.v.c. bus; either will effect a cure, the latter being the better method.

A microphone can be temporarily borrowed for the emergency from the nearest telephone. Two output transformers, of the type designed to match a pentode tube to dynamic speaker voice coil, can be connected with their corresponding windings in series (primaries in series; secondaries in series) and used backwards as a single, microphone transformer to feed directly into a pentode tube used as a class A modulator (Heising). It will be necessary to talk loudly and directly into the microphone, which incidentally should be fed from a separate 6 volt battery in order to avoid vibrator hash in the speech.

Home stations often can be kept on the air after a power failure by utilizing an auto radio and battery, feeding plate voltage to either the oscillator or exciter of your regular transmitter from the vibrator pack and deriving heater voltage direct from the battery. The oscillator or exciter should be originally designed with this in mind.

Any amateurs in the vicinity possessing battery-powered or auto-installed 2½ meter gear can locate their equipment at strategic points and take care of short-haul traffic.

Amateurs not in the immediate disaster area usually can be of most service by doing lots of listening, and little or no transmitting until called upon.


8005
816
815
931
826

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MUNICATION'S MOST EXACTING APPLICATIONS

Whether you want a Midget tube for UHF experimentation on an amateur rig or a giant, air-cooled tube for commercial broadcasting, RCA makes it. The RCA line is complete—and, above all, each tube represents the utmost in long, trouble-free performance as proved time and again on radio's toughest jobs. When you invest in an RCA tube you invest in experience—not experiments!

TRANSMITTING TUBES

No.	Type	Max. Input	Net Price
801-A	Triode	42 Watts	$3.45
802	Pentode	33 Watts*	3.50
803	Pentode	350 Watts	28.50
804	Pentode	150 Watts*	15.00
805	Triode	315 Watts	13.50
806	Triode	1000 Watts*	22.00
807	Beam	75 Watts*	3.50
808	Triode	200 Watts	7.75
809	Triode	100 Watts*	2.50
810	Triode	620 Watts*	13.50
811	Triode	225 Watts*	3.50
812	Triode	225 Watts*	3.50
813	Beam	360 Watts	22.00
815	UHF Twin Beam	75 Watts*	4.50
825	Inductive Output	100 Watts	34.50
826	UHF Triode	125 Watts	19.00
828	Beam	270 Watts*	17.50
833-A	Triode	2000 Watts*	85.00
834	UHF Triode	125 Watts	12.50
1623	Triode	100 Watts*	2.50
1624	Beam	54 Watts	3.50
1628	UHF Triode	50 Watts	32.00
8000	Triode	620 Watts*	13.50
8001	Beam	300 Watts	27.50
8003	Triode	330 Watts	12.00
8005	De Luxe Triode	300 Watts*	7.00

*ICAS Rating

HALF-WAVE MERCURY-VAPOR RECTIFIERS

No.	Max. Peak Inverse Voltage	Max. Average Plate Current	Net Price
816	5,000 Volts	.125 a.	$1.00
866-A/866	10,000 Volts	0.25 a.	1.50
872	7,500 Volts	1.25 a.	9.00
872-A	10,000 Volts	1.25 a.	11.00

TELEVISION TUBES

No.	Description	Net Price
3AP4/906P4	3" Kinescope	$13.75
5AP4/1805P4	5" Kinescope (Short Bulb)	22.00
5BP4/1802P4	5" Kinescope	22.00
1847	Amateur Iconoscope	24.50

CATHODE-RAY TUBES

No.	Screen	Net Price
3AP1/906P1	3" Green Phosphor	$13.50
902	2" Green Phosphor	7.50
913	1" Green Phosphor	4.00

UHF ACORN TUBES

No.	Description	Net Price
954	Pentode. Amplifier, Detector	$5.00
955	Triode. Detector, Oscillator	3.00
956	Pentode. Super-Control Amp.	5.00
957	Triode. Low-drain filament	3.00
958	Triode. Low-drain filament	3.00
959	Pentode. Low-drain filament	5.00

UHF MIDGETS

No.	Description	Net Price
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9002	Triode. Detector, Oscillator	2.00
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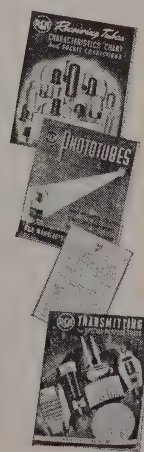


SERVICE OF THE RADIO CORPORATION OF AMERICA • *For Canadian prices write to RCA Victor Company Limited, Montreal*

Article 24

542 1. No provision of these Regulations shall prevent a mobile station in distress from using any means available to it for drawing attention, signalling its position, and obtaining help.

* * * *

548 3. (2) Aircraft. Any aircraft in distress must transmit the distress call on the watching-wave of the land or mobile stations capable of helping it; when the call is addressed to stations of the maritime service, the waves to be used are the distress-wave or watching-wave of these stations.

549 § 4. (1) In radiotelegraphy, the distress signal shall consist of the group • • • — — — • • • transmitted as one signal, in which the dashes must be emphasized so as to be distinguished clearly from the dots.

550 In radiotelephony, the distress signal shall consist of the spoken expression Mayday (corresponding to the French pronunciation of the expression "m'aider").

551 (2) These distress signals shall announce that the ship, aircraft, or any other vehicle which sends the distress signal is threatened by serious and imminent danger and requests immediate assistance.

* * * *

555 § 5. (4) This call shall have absolute priority over other transmissions. All stations hearing it must immediately cease all transmission capable of interfering with the distress traffic and must listen on the wave used for the distress call. This call must not be sent to any particular station and shall not require an acknowledgment of receipt.

556 § 6. (1) The distress call must be followed as soon as possible by the distress message. This message shall include the distress call followed by the name of the ship, aircraft, or the vehicle in distress, information regarding the position of the latter, the nature of the distress and the nature of the help requested, and any other further information which might facilitate this assistance.

557 (2) When, in its distress message, an aircraft is unable to signal its position, it shall endeavor after the transmission of the incomplete message to send its call signal long enough so that the radio direction-finding stations may determine its position.

558 § 7. (1) As a general rule, a ship or aircraft at sea shall signal its position in latitude and longitude (Greenwich) using figures, for the degrees and minutes, accompanied by one of the words North or South and one of the words East or West. A period shall separate the degrees from the minutes. In some cases, the true bearings and the distance in nautical miles from some known geographical point may be given.

* * * *

560 (3) As a general rule, an aircraft flying over land shall signal its position by the name of the nearest locality, its approximate distance from this point, accompanied, according to the case, by one of the words North, South, East, or West, or, in some cases, words indicating intermediate directions.

561 § 8. The distress call and message shall be sent only by order of the master or person responsible for the ship, aircraft, or other vehicle carrying the mobile station.

* * * *

569 § 11. (1) Stations of the mobile service which receive a distress message from a mobile station which is unquestionably in their vicinity, must acknowledge receipt thereof at once (see Nos. 587, 588, and 589). If the distress call has not been preceded by an auto-alarm signal, these stations may transmit this auto-alarm signal with the authorization of the authority responsible for the station (for mobile stations, see No. 276), taking care not to interfere with the transmission of the acknowledgment of the receipt of said message by other stations.

570 (2) Stations of the mobile service which receive a distress message from a mobile station which unquestionably is not in their vicinity, must wait a short period of time before acknowledging receipt thereof, in order to make it possible for stations nearer to the mobile station in distress to answer and acknowledge receipt without interference.

* * * *

573 § 14. The control of distress traffic shall devolve upon the mobile station in distress or upon the mobile station which, by application of the provisions of No. 567, has sent the distress call. These stations may delegate the control of the distress traffic to another station.

* * * *

604 § 22. (2) In radiotelephony the urgent signal shall consist of three transmissions of the expression PAN (corresponding to the French pronunciation of the word "panne"); it shall be transmitted before the call.

605 (3) The urgent signal shall indicate that the calling station has a very urgent message to transmit concerning the safety of a ship, an aircraft, or another vehicle, or concerning the safety of some person on board or sighted from on board.

606 (4) In the aeronautical service, the urgent signal PAN shall be used in radiotelegraphy and in radiotelephony to indicate that the aircraft transmitting it is in trouble and is

forced to land, but that it is not in need of immediate help. This signal should, so far as possible, be followed by a message giving additional information.

607 (5) The urgent signal shall have priority over all other communications, except distress communications, and all mobile or land stations hearing it must take care not to interfere with the transmission of the message which follows the urgent signal.

608 (6) In case the urgent signal is used by a mobile station, this signal must, as a general rule, subject to the provisions of No. 606, be addressed to a definite station.

* * * *

612 § 25. (1) The urgent signal may be transmitted only with the authorization of the master or of the person responsible for the ship, aircraft, or any other vehicle carrying the mobile station.

613 (2) In the case of a land station, the urgent signal may be transmitted only with the approval of the responsible authority.

* * * *

615 § 26. (1) In radiotelegraphy, the safety signal shall consist of the group TTT, transmitted three times, with the letters of each group, as well as the consecutive groups, well separated. This signal shall be followed by the word DE and three transmissions of the call signal of the station sending it. It announces that this station is about to transmit a message concerning the safety of navigation or giving important meteorological warnings.

616 (2) In radiotelephony, the word Security (corresponding to the French pronunciation of the word "sécurité") repeated three times, shall be used as the safety signal.

* * * *

619 § 28. (2) All stations hearing the safety signal must continue listening on the wave on which the safety signal has been sent until the message so announced has been completed; they must moreover keep silence on all waves likely to interfere with the message.

Article 26

Order of Priority of Communications in the Mobile Service

653 The order of priority of radio communications in the mobile service shall be as follows:

1. Distress calls, distress messages, and distress traffic;
2. Communications preceded by an urgent signal;
3. Communications preceded by a safety signal;
4. Communications relative to radio direction-finding bearings;
5. Government radiotelegrams for which priority right has not been waived;
6. All other communications.

R-S-T REPORTING SYSTEM

Readability

1. Unreadable.
2. Barely Readable—Occasional Words Distinguishable.
3. Readable with Considerable Difficulty.
4. Readable with Practically No Difficulty.
5. Perfectly Readable.

Signal Strength

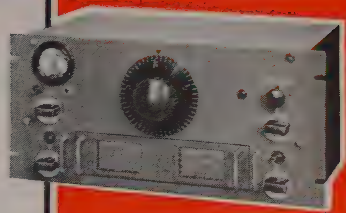
1. Faint—Signals Barely Perceptible.
2. Very Weak Signals.
3. Weak Signals.
4. Fair Signals.
5. Fairly Good Signals.
6. Good Signals.
7. Moderately Strong Signals.
8. Strong Signals.
9. Extremely Strong Signals.

Tone

1. Extremely Rough, Hissing Note.
2. Very Rough A.C. Note—No Trace of Musicality.
3. Rough, Low-Pitched A.C. Note—Slightly Musical.
4. Rather Rough A.C. Note—Moderately Musical.
5. Musically Modulated Note.
6. Modulated Note—Slight Trace of Whistle.
7. Near D.C. Note—Smooth Ripple.
8. Good D.C. Note—Just Trace of Ripple.
9. Purest D.C. Note.

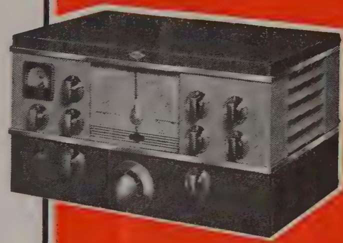
If the Note Appears to Be Crystal Controlled, Simply Add an X after the Appropriate Number.

WHAT IT TAKES



HRO

National Receivers give maximum performance in every price range, at all frequencies. There is the HRO, designed for amateurs but used for tough jobs everywhere, in the Wireless Room of the British Admiralty, for instance. There is the little SW-3-U, newly redesigned, but retaining the sterling qualities that have made its forerunner, the SW-3, a favorite for eleven years.



NC 200

There is the NC-200 which has made a reputation for itself in one short year. There is the brand new NC-45 which brings top performance in its price class.

For ultra high frequencies, the One-Ten with coverage from one to ten meters, is still *the* receiver for work at 112 MC to 224 MC. The deluxe NHU brings communication receiver performance to the 5-meter band as well as 10 and 20 meters.

NC 100

These, with the NC-100 series, make up the roll call of National Receivers for amateur use. Different models in each of the above types make up a total of more than twenty receivers to choose from. Regardless of your needs, you will find a National Receiver to fill them.

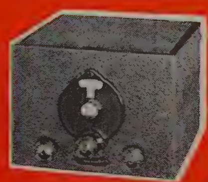
NC 45

NATIONAL COMPANY, INC.

MALDEN, MASSACHUSETTS

NHU

ONE-TEN



SW-3U

Abbreviation	Question	Answer
QRA	What is the name of your station?	The name of my station is ....
QRB	How far approximately are you from my station?	The approximate distance between our stations is .... nautical miles (or .... kilometers).
QRC	What company (or Government Administration) settles the accounts for your station?	The accounts for my station are settled by the .... company (or by the Government Administration of ....).
QRD	Where are you bound and where are you from?	I am bound for .... from ....
QRG	Will you tell me my exact frequency (wavelength) in kc/s (or m)?	Your exact frequency (wavelength) is .... kc/s (or .... m).
QRH	Does my frequency (wavelength) vary?	Your frequency (wavelength) varies.
QRI	Is my note good?	Your note varies.
QRJ	Do you receive me badly? Are my signals weak?	I cannot receive you. Your signals are too weak.
QRK	Do you receive me well? Are my signals good?	I receive you well. Your signals are good.
QRL	Are you busy?	I am busy (or I am busy with ......). Please do not interfere.
QRM	Are you being interfered with?	I am being interfered with.
QRN	Are you troubled by atmospherics?	I am troubled by atmospherics.
QRO	Shall I increase power?	Increase power.
QRP	Shall I decrease power?	Decrease power.
QRQ	Shall I send faster?	Send faster (........ words per minute).
QRR	....	*Amateur "SOS" or distress call (U.S.A.). Use only in serious emergency.*
QRS	Shall I send more slowly?	Send more slowly (........ words per minute).
QRT	Shall I stop sending?	Stop sending.
QRU	Have you anything for me?	I have nothing for you.
QRV	Are you ready?	I am ready.
QRW	Shall I tell .... that you are calling him on .... kc/s (or...... m)?	Please tell .... that I am calling him on .... kc/s (or .... m).
QRX	Shall I wait? When will you call me again?	Wait (or wait until I have finished communicating with ....) I will call you at .... o'clock (or immediately).
QRY	What is my turn?	Your turn is No. .... (or according to any other method of arranging it).
QRZ	Who is calling me?	You are being called by ....
QSA	What is the strength of my signals (1 to 5)?	The strength of your signals is .... (1 to 5).
QSB	Does the strength of my signals vary?	The strength of your signals varies.
QSD	Is my keying correct; are my signals distinct?	Your keying is incorrect; your signals are bad.

Abbreviation	Question	Answer
QSG	Shall I send .... telegrams (or one telegram) at a time?	Send .... telegrams (or one telegram) at a time.
QSJ	What is the charge per word .... including your internal telegraph charge?	The charge per word for .... is .... francs, including my internal telegraph charge.
QSK	Shall I continue with the transmission of all my traffic, I can hear you through my signals?	Continue with the transmission of all your traffic, I will interrupt you if necessary.
QSL	Can you give me acknowledgment of receipt?	I give you acknowledgment of receipt.
QSM	Shall I repeat the last telegram I sent you?	Repeat the last telegram you have sent me.
QSO	Can you communicate with .... direct (or through the medium of ....)?	I can communicate with .... direct (or through the medium of ....).
QSP	Will you retransmit to .... free of charge?	I will retransmit to .... free of charge.
QSR	Has the distress call received from .... been cleared?	The distress call received from .... has been cleared by ....
QSU	Shall I send (or reply) on .... kc/s (or m) and/or on waves of Type A1, A2, A3, or B?	Send (or reply) on .... kc/s (or .... m) and/or on waves of Type A1, A2, A3, or B.
QSV	Shall I send a series of VVV ....?	Send a series of VVV ....
QSW	Will you send on .... kc/s (or .... m) and/or on waves of Type A1, A2, A3 or B?	I am going to send (or I will send) on .... kc/s (or .... m) and/or on waves of Type A1, A2, A3 or B.
QSX	Will you listen for .... (call sign) on .... kc/s (or .... m)?	I am listening for .... (call sign) on .... kc/s (or .... m).
QSY	Shall I change to transmission on .... kc/s (or .... m) without changing the type of wave? or	Change to transmission on .... kc/s (or .... m) without changing the type of wave.
	Shall I change to transmission on another wave?	Change to transmission on another wave.
QSZ	Shall I send each word or group twice?	Send each word or group twice.
QTA	Shall I cancel telegram No. .... as if it had not been sent?	Cancel telegram No. .... as if it had not been sent.
QTB	Do you agree with my number of words?	I do not agree with your number of words; I will repeat the first letter of each word and the first figure of each number.
QTC	How many telegrams have you to send?	I have .... telegrams for you (or for ....).
QTE	What is my true bearing in relation to you? or	Your true bearing in relation to me is .... degrees or
	What is my true bearing in relation to .... (call sign)? or	Your true bearing in relation to ...... (call sign) is .... degrees at .... (time) or
	What is the true bearing of .... (call sign) in relation to .... (call sign)?	The true bearing of .... (call sign) in relation to .... (call sign) is .... degrees at .... (time).

EMERGENCY

The door is locked on the old ham shack. Emergency. Uncle Sam calling. A **national** emergency this time... unlimited. It took a lot of black headlines and some strong legislation before most Americans were awakened, stirred to action. But not the amateur radio operator. Like the Navy, he is always ready. Emergency is, in fact, his middle name.

Today no branch of the Service is without its amateur radio operators. Hams have been called to active duty in the U. S. Navy, the Merchant Marine, the Army Signal Corps, the FBI. Still other amateurs are standing by, at their posts, awaiting special orders. Indeed, when the history of this war is recorded, the American ham—his readiness, willingness and fitness to help get on with the job of building a strong national defense through an efficient armed force—must rank high among citations for distinguished service.

The qualified amateur, answering the call to the colors, is already trained. He knows the Code. He knows too, because his middle name is Emergency, how important in radio

why that's a ham's middle name!

communication is dependable equipment. And so C-D capacitors, having served so many amateur radio operators so faithfully for so many years, today are serving Uncle Sam. Cornell-Dubilier experience—the longest experience back of capacitors—is responsible for the amazing dependability of C-Ds. Next time you need capacitors, specify Cornell-Dubilier. You can depend on them more than ever for **they're in the Army now!** Address Cornell-Dubilier Electric Corp., 1013 Hamilton Blvd., South Plainfield, N. J.

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ELECTRIC CORPORATION

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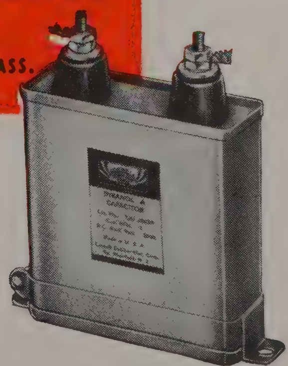

THE MOST DEPENDABLE units the amateur can choose are Cornell-Dubilier type TJU Dykanol Capacitors. These are the capacitors practically every broadcast and government station in the world uses with signal success. Safely rated, compact and lightweight, they are standard equipment with thousands of amateurs.

CORNELL-DUBILIER CD CAPACITORS

Abbreviation	Question	Answer
QTF	Will you give me the position of my station according to the bearings taken by the direction-finding stations which you control?	The position of your station according to the bearings taken by the direction-finding stations which I control is .... latitude ....longtitude.
QTG	Will you send your call sign for fifty seconds followed by a dash of ten seconds on .... kc/s (or .... m) in order that I may take your bearing?	I will send my call sign for fifty seconds followed by a dash of ten seconds on .... kc/s (or .... m) in order that you may take my bearing.
QTH	What is your position in latitude and longitude (or by any other way of showing it)?	My position is........latitude........ longitude (or by any other way of showing it).
QTI	What is your true course?	My true course is ....degrees.
QTJ	What is your speed?	My speed is .... knots (or .... kilometers) per hour.
QTM	Send radioelectric signals and submarine sound signals to enable me to fix my bearing and my distance.	I will send radioelectric signals and submarine sound signals to enable you to fix your bearing and your distance.
QTO	Have you left dock (or port)?	I have just left dock (or port).
QTP	Are you going to enter dock (or port)?	I am going to enter dock (or port).
QTQ	Can you communicate with my station by means of the International Code of Signals?	I am going to communicate with your station by means of the International Code of Signals.
QTR	What is the exact time?	The exact time is ....
QTU	What are the hours during which your station is open?	My station is open from .... to ....
QUA	Have you news of .... (call sign of the mobile station)?	Here is news of .... (call sign of the mobile station).
QUB	Can you give me in this order, information concerning: visibility, height of clouds, ground wind for .... (place of observation)?	Here is the information requested ....
QUC	What is the last message received by you from ...... (call sign of the mobile station)?	The last message received by me from .... (call sign of the mobile station) is ....
QUD	Have you received the urgency signal sent by .... (call sign of the mobile station)?	I have received the urgency signal sent by .... (call sign of the mobile station) at .... (time).
QUF	Have you received the distress signal sent by .... (call sign of the mobile station)?	I have received the distress signal sent by .... (call sign of the mobile station) at .... (time).
QUG	Are you being forced to alight in the sea (or to land)?	I am forced to alight (or land) at .... (place).
QUH	Will you indicate the present barometric pressure at sea level?	The present barometric pressure at sea level is .... (units).
QUJ	Will you indicate the true course for me to follow, with no wind, to make for you?	The true course for you to follow, with no wind, to make for me is .... degrees at .... (time).

ABBREVIATIONS COMMONLY USED BY AMATEURS

ABT	About	MO	More
AGN	Again	MSG	Message
AHD	Ahead	MT	Empty
AHR	Another	N	No
ANI	Any	ND	Nothing Doing
APRX	Approximate-Approximately	NG	No Good
BC	Broadcast	NIL	Nothing
BD	Bad	NM	No More
B4	Before	NR	Number
BK	Break	NW	Now
BN	Been	OB	Old Boy
BND	Band	OL	Old Lady
BCUZ	Because	OM	Old Man
BTWN	Between	OP	Operator
BIZ	Business	OT	Old Top—Timer
C	See, Yes	OW	Old Woman
CLR	Clear	PLS	Please
CN	Can	PSE	Please
CNT	Can't	PX	Press
CK	Check	R	OK
CKT	Circuit	RCD	Received
CMG	Coming	RCVR	Receiver
CUD	Could	RI	Radio Inspector
CW	Continuous Wave	SA	Say
CUL	See You Later	SEZ	Says
CUAGN	See You Again	SM	Some
DE	From	SW	Short-wave
DA	Day	SIG	Signal
DNT	Don't	SKED	Schedule
DINT	Did Not	TFC	Traffic
DH	Deadhead	TMW	Tomorrow
DX	Long Distance	TR	There
ES	And	TT	That
EZ	Easy	TK	Take
FB	Fine Business	TKS	Thanks
FM	From	TNK	Think
FR	For	TNX	Thanks
FRQ	Frequency	U	You
GA	Go Ahead	UD	You Would
GB	Good-Bye	UL	You Will
GM	Good Morning	UR	Your
GN	Good Night	VT	Vacuum Tube
GG	Going	VY	Very
GT	Got—Get	WA	Word After
GND	Ground	WB	Word Before
HA(HI)	Laughter	WD	Would
HM	Him	WF	Word Following
HR	Here—Hear	WK	Work
HV	Have	WL	Will—Would
HW	How	WN	When
IC	I See	WT	What
ICW	Interrupted Continuous Wave	WX	Weather
K	Go Ahead	X	Interference
LID	Poor Operator	XMTR	Transmitter
LIL	Little	XTAL	Crystal
LFT	Left	YF	Wife
LST	Last—Listen	YL	Young Lady
LTR	Letter	YR	Your
MG	Motor Generator	30	Finish—End
MI	My	73	Best Regards
MK	Make	88	Love and Kisses


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40 METERS
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TYPE SOC100
80-160 METERS
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Guaranteed Accuracy
SEE CATALOG G-12
BLILEY ELECTRIC CO.

ARTIFICIAL RESPIRATION

By the Prone Pressure Method

(Illustrations Courtesy of National Safety Council, Chicago)

The following is the accepted, standardized technique of "How to Give Artificial Respiration by the Prone Pressure Method," agreed upon by a special committee of national organizations and persons appointed by the United States Public Health Service of the Treasury Department.

The Prone Pressure Method of artificial respiration described in these rules should be used in cases of suspended respiration from all causes—drowning, *electric shock,* carbon monoxide poisoning, injuries, etc. Delay of even one minute in the application of the method may lose a life. Follow the instructions even if the patient appears dead. Continue artificial respiration until natural breathing is restored or until a physician declares rigor mortis (stiffening of the body) has set in. Success *has come* after three and one half hours of effort.

Learn this method now. Don't wait for an accident. Practice on a friend. Let him practice on you.

1. Lay the patient on his belly, one arm extended directly overhead, the other arm bent at elbow and with the face turned outward and resting on hand or forearm so that the nose and mouth are free for breathing. (See figure 1.)

2. Kneel straddling the patient's thighs with your knees placed at such a distance from the hip bones as will allow you to assume the position shown in figure 1.

Place the palms of the hands on the small of the back with fingers resting on the ribs, the little finger just touching the lowest rib, with the thumb and fingers in a natural position, and the tips of the fingers just out of sight. (See figure 1.)

3. With arms held straight, swing forward slowly so that the weight of your body is gradually brought to bear upon the patient. The shoulder should be directly over the heel of the hand at the end of the forward swing. (See figure 2.) Do not bend your elbows. This operation should take about two seconds.

4. Now immediately swing backward so as to completely remove the pressure, thus returning to the position in figure 3.

5. After two seconds, swing forward again. Thus repeat deliberately twelve to fifteen times a minute the double movement of compression and release, a complete respiration in four or five seconds.

6. Continue artificial respiration without interruption until natural breathing is restored if necessary, four hours or longer, or until a physician declares the patient is dead.

7. As soon as this artificial respiration has been started and while it is being continued, an assistant should loosen any tight clothing about the patient's neck, chest, or waist. Keep the patient warm. Do not give any liquids whatever by mouth until the patient is fully conscious.

8. To avoid strain on the heart when the patient revives, he should be kept lying down and not allowed to stand or sit up. If the doctor has not arrived by the time the patient has revived, he should be given some stimulant, such as one teaspoonful of aromatic spirits of ammonia in a small glass of water or a hot drink of coffee or tea, etc. The patient should be kept warm.

9. Resuscitation should be carried on at the nearest possible point to where the patient received his injuries. He should not be moved from this point until he is breathing normally of his own volition and then moved only in a lying position. Should it be necessary, due to extreme weather conditions, etc., to move the patient before he is breathing normally, resuscitation should be carried on during the time he is being moved, if practicable.

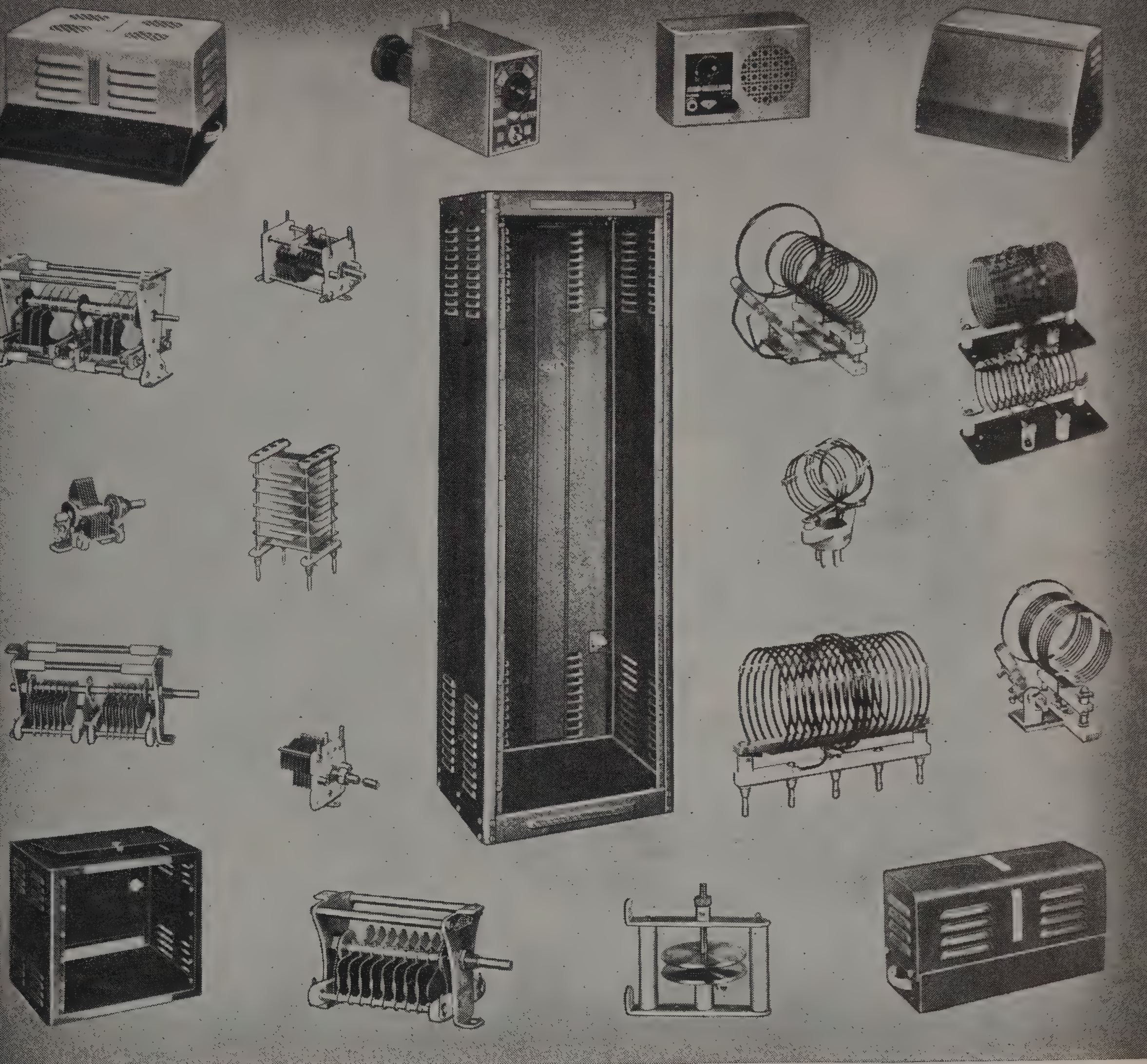
10. A brief return of natural respiration is not a certain indication for stopping the resuscitation. Not infrequently the patient, after a temporary recovery of respiration, stops breathing again. The patient must be watched and if natural breathing stops, artificial respiration should be resumed at once.

11. In carrying out resuscitation it may be necessary to change the operator. This change must be made without losing the rhythm of respiration. By this procedure no confusion results at the time of change of operator and a regular rhythm is kept up.

FIGURE 1

FIGURE 2

FIGURE 3

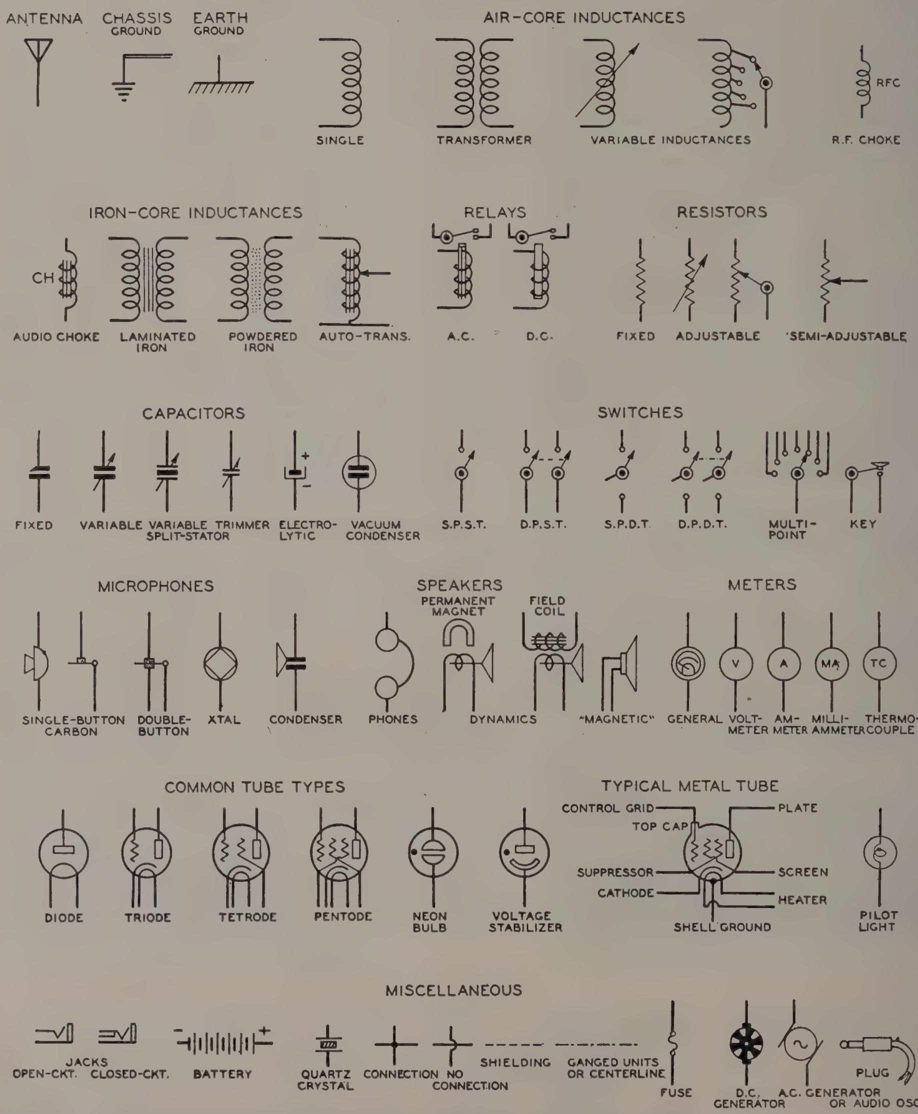


FRACTIONAL-DECIMAL EQUIVALENTS

A time-saving table is given for fractional-decimal conversion. Many of the commonly used fractions and their decimal equivalents are shown. Others can be calculated by dividing the numerator by the denominator.

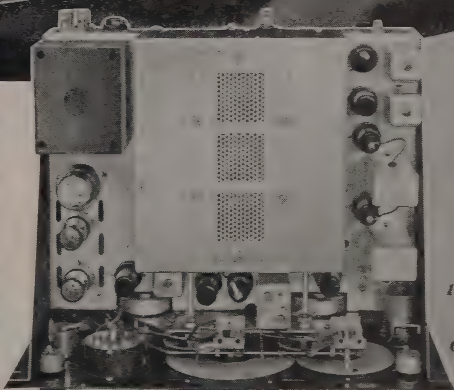
1/64 = .0165	1/8 = .125	7/16 = .4375	3/4 = .750
1/32 = .0312	3/16 = .1875	1/2 = .500	13/16 = .8125
3/64 = .0468	1/4 = .250	9/16 = .5625	7/8 = .875
1/16 = .0625	5/16 = .3125	5/8 = .625	15/16 = .9375
3/32 = .0936	3/8 = .3750	11/16 = .6825	

RADIO SYMBOLS USED IN CIRCUIT DIAGRAMS

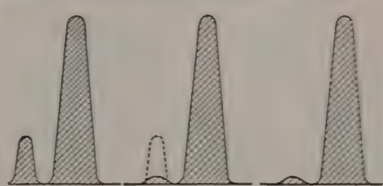


Precision Performance!

The New 1941 Super Skyrider Model SX-28 delivers precision performance! 15 tubes. 2 stages pre-selection. Calibrated bandspread inertia controlled. Micrometer scale tuning inertia controlled. Tone and AC on-off. Beat frequency oscillator. AF gain. RF gain. Crystal phasing. Adjustable noise limiter. Send-receive switch. AVC-BFO switch. Bass boost switch. Phono jack. 80/-40/20/10 meter amateur bands calibrated. Wide angle "S" meter. Band pass audio filter. Improved signal to image and noise ratio. Push-pull high fidelity audio output. 6 step wide range variable selectivity. Improved headphone output. Super Skyrider, Model SX-28 with crystal and tubes, $179.50.



Interior View SX-28 Chassis



With selectivity switch in XTAL sharp position identify the weaker amplitude — Tune Receiver to the weaker.

Adjust phasing control carefully until this weaker amplitude is reduced to a minimum.

Retune Receiver to the stronger amplitude and then adjust pitch control until you get note most pleasing to copy.

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CHICAGO, U. S. A.

USED BY 33 GOVERNMENTS
SOLD IN 89 COUNTRIES

Class B
Amateur License Examinations

The Federal Communications Commission has prepared a reservoir of some several hundred questions for the amateur examination. After you have successfully passed your code test, a group of these questions will be selected from the reservoir and you must make a grade of 75 per cent or higher; otherwise you must wait at least two months from the date of the examination and attempt the examination again.

The questions are changed from time to time to keep pace with revisions in the regulations and with technical progress. However, the applicant can be sure of receiving one question from each of the following ten general classes: Transmitter Theory; Transmitter Practice; Radiotelephony; Power Supplies; Frequency Measurement; Treaty and Laws; F.C.C. Regulations, Bands; F.C.C. Regulations, Part I; F.C.C. Regulations, Part II; Penalties.

The following paraphrased questions are representative of the scope of the *technical* questions contained in the amateur radio operator's license examination.

No questions on laws, penalties, etc. are included, because it would be impossible to give suitable answers with the assurance that they would be correct at the time the applicant took his examination. Therefore the applicant should study carefully the latest data pertaining to laws and penalties as obtained from the Government Printing Office, and be able to answer any question pertaining thereto.

The answers given here to the technical questions are not *necessarily* the *only* correct answers; neither do we guarantee that all the answers given here would command a "100% correct" grading from the F.C.C. You may be assured, however, that if you can answer all of the questions given here and have a pretty good idea as to why each answer is correct, you need have no fear of failing to make a passing grade on the technical portion of the examination.

If you have difficulty in understanding why the answer to a particular question is correct, more study of the theory applying to that question is indicated.

The actual questions as given in the examination will appear in the short answer form such as multiple choice or simple diagrams and computations, etc.

> The FCC cannot answer inquiries from candidates who have taken an examination as to what grade they made, or what was the matter with their answers if they did not pass. The large number of candidates makes this impossible.

Class B Study Guide

1. Name the basic units of electrical resistance, inductance, capacitance, current, electromotive force or potential difference, power, energy, quantity, magnetomotive force, and frequency.

Basic Units: Resistance—ohm; Inductance—henry; Capacitance—farad; Current—ampere; Electromotive Force or Potential Difference—volt; Power—watt; Energy—joule; Quantity—coulomb; Magnetomotive Force—gilbert; Frequency—cycles per second.

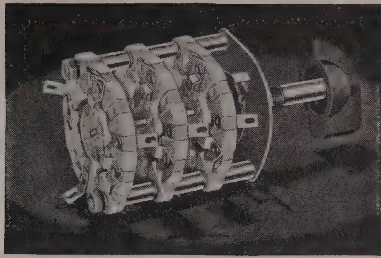
2. Name the instruments normally used to measure:
(a) electric current
(b) potential difference
(c) power
(d) resistance
(e) frequency

(a) ammeter
(b) voltmeter
(c) wattmeter
(d) ohmmeter
(e) frequency meter or wave meter.

3. How may plate power input of an amplifier be determined when the plate voltage and plate current are known?

Plate power input may be determined by applying the power formula, $P=EI$ (or $W=EI$), which means that the power equals the product of the voltage in volts and the current in amperes.

Centralab
CRL
The Quality Line



SELECTOR SWITCH
Available in an almost infinite variety of combinations . . . in bakelite or steatite . . . in single or multiple gang . . . from two to eleven positions on any one switch . . . also available for use in amateur transmitters.

Hams, Servicemen, Experimenters and Manufacturers appreciate the utter dependability of Centralab products. Since 1922 more than a hundred million radio parts bespeak the universal acceptance accorded Centralab products. Send for catalog if your jobber cannot supply you.

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Div. of Globe-Union Inc.
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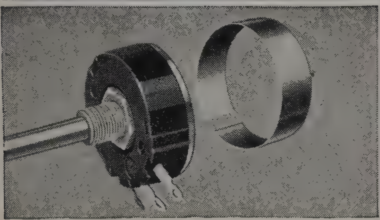
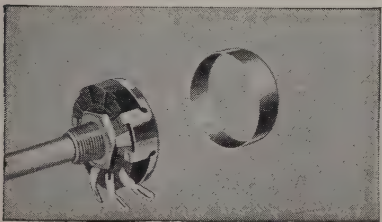
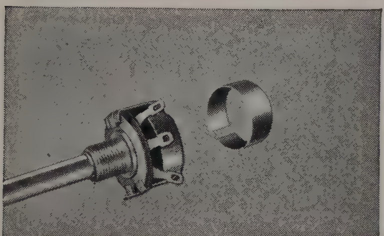


STANDARD RADIOHM
Wall type resistor. Exclusive non-rubbing contact band. 1⅜" diameter x 9/16" deep. Available single, twin or triple, plain or tapped . . . with S.P.S.T., D.P.S.T. or S.P.D.T.

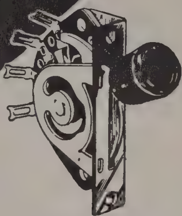


MIDGET RADIOHM
Companion to "standard" . . . small size but large control efficiency. Available single, dual or triple . . . plain or one, two or three taps . . . with S.P.S.T., S.P.D.T., or D.P.S.T. Moulded bakelite case, 1⅛" diameter, ¼" metal shaft 3⅜" long.

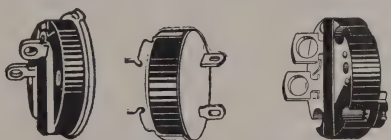


ELF RADIOHM
Smaller but also features the long, straight resistor strip. Available plain or tapped with S.P.S.T. switch . . . with or without dummy lug. Bakelite case 57/64" diameter, 17/32" deep (less switch) 25/32" deep with switch.

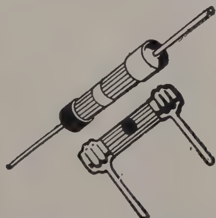
WIRE WOUND RADIOHMS
In values from 2 to 10,000 ohms . . . insulated construction . . . 3 watts . . . universal shaft for all replacements . . . regular Radiohm switch covers may be attached . . . in linear curve only . . .



LEVER ACTION SWITCH
Used singly or in groups . . . for broadcasting, receiving, public address, test instruments and industrial uses. Available in any one of ten different combinations including positive and spring return action.

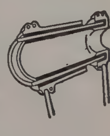


ATTACHABLE SWITCH COVERS
For standard and wirewound resistors (Radiohms) as well as Midget and Elf Radiohms . . . S.P.S.T. . . . S.P.D.T. . . . D.P.S.T . . . four point . . . S.P.D.T. (operates at clockwise position) and S.P.S.T. with Dummy Lug.



AXIAL LEAD RESISTORS
Body is insulated by inert ceramic jacket . . . proof against vibration and humidity . . . will withstand five times rated load without permanent change. In two sizes . . . RMA coded . . . ½ watt at ⅛" x ⅝" and 1 watt at ¼" x 1" . . . Also supplied in conventional RADIAL LEAD Style . . . ½ watt — 1 watt or 2 watt.

CERAMIC TRIMMER CONDENSER
where greater stability than ordinary types is required. Supplied with neg. temp. coefficient of .006 MMF/MMF/C°. With or without mounting brackets.



CERAMIC CAPACITOR
Small "special purpose" for h.f. circuits where temperature compensation, low power factor, or absolute permanence are important. 1000 V.D.C. leakage resistance more than 10,000 meg. Power factor less than 1%.

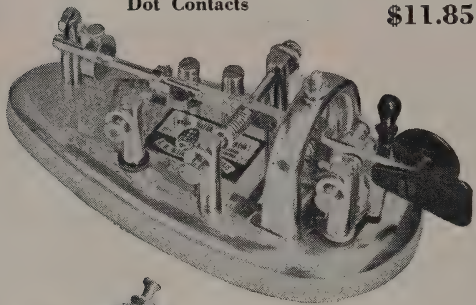
SEND FOR CATALOG NO. 23

RADIO TELEGRAPH APPARATUS *Manufactured by* WORLD'S CHAMPION RADIO TELEGRAPHER

NEW SUPER STREAM-SPEED

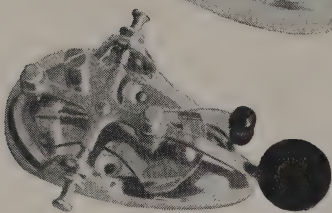
PLATINUM
Dot Contacts

S-600 PC
$11.85



Into this gorgeous speed key has gone Mac's 30 years operating experience supplemented by the finest engineering ability in the radio-telegraph industry . . . with their combined efforts coordinated under the styling genius of one of America's outstanding design artists. See it! Handle it! You'll have to own it! Combining beauty and utility in a most striking fashion, this radically new, semi-automatic key is the last word in operating ease. Fast, rhythmical Morse is a real pleasure with this key.

- **Streamlined base of special dense alloy. Wt. 4 lbs.**
- **Tear-drop shaped base makes it immovable on table.**
- **Heavily chromed with bluish tinge to prevent glare.**
- **Swedish blued steel mainspring and U spring.**
- **Bronze bearing screws.**
- **Bronze alloy pigtail.**
- **Bakelite insulation throughout.**
- **Molded plastic dot paddle and dash button.**



STREAM KEY DE LUXE MODEL

No. 300 at $3.45 net.
No. 200 at $2.25 net.
No. S-100 at $1.65 net.

Beautiful tear-drop streamlined base with heavy bluish tinged chrome. All parts chromed. Finely balanced key lever. 3/16″ contacts, designed expressly for these keys. These pretty Streamkeys have a "feel" that makes good Morse easy for any operator.

PRACTICE SET

CATALOG NO. S-710—$10.50 Net
Complete with tube

PROFESSIONAL MODEL, same key but base "black wrinkled." CATALOG No. 200

AMATEUR MODEL, rich black polystyrene base, same lever, and large contacts, circuit closing switch, bronze bearing screws. A beautiful and superbly balanced key at an absurdly low price because of enormous volume. CATALOG No. S-100

PROFESSIONAL MODEL, MAC KEY

Designed to conform with United States Navy specifications for "speed key." It is just what its name implies: A fine Professional Operator's model Mac Key. Base 3¾″ x 6½″ x ¾″ thickness. Beautifully black wrinkled over Parkerized base casting. Carefully designed super-structure, similarly finished.

Chromed parts, circuit closer, bakelite insulation, 3/16″ silver contacts. A key that will thrill any radio or telegraph operator.

TELEGRAPHER'S MODEL CHROMED YOKE, CORD. CP500 @ $9.50

MODEL No. P500
$7.50
NET TO THE OPERATOR

T. R. McElroy

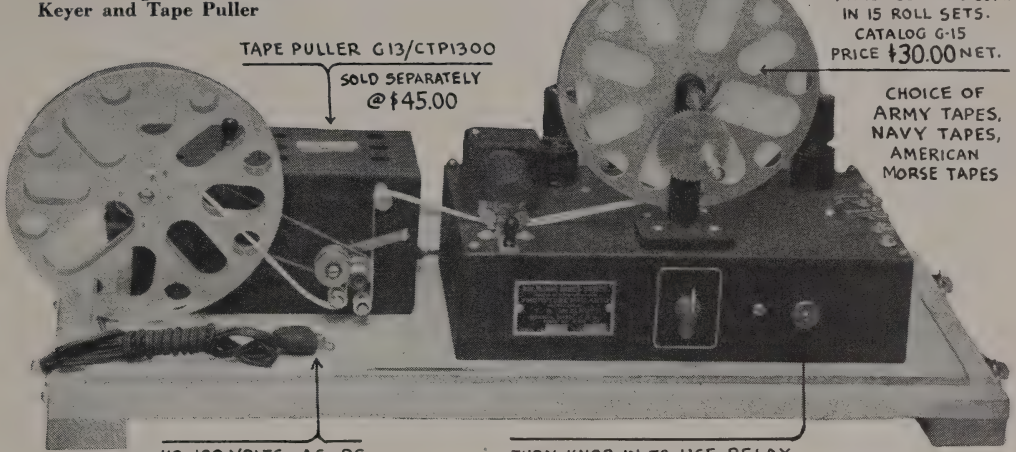
MAC

82 Brookline Ave., Boston, Mass., U. S. A.

RADIO TELEGRAPH APPARATUS *Manufactured by* WORLD'S CHAMPION RADIO TELEGRAPHER

PRICE $90.00 Net

G-813 is complete as shown Keyer and Tape Puller



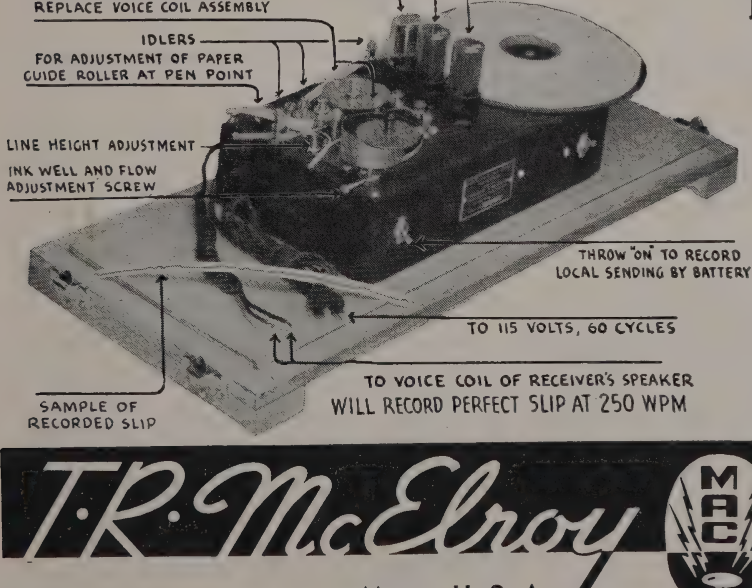
McELROY MODEL G-813 RADIO CODE KEYING UNIT

Designed for radio operator training schools of the United States Army.

Dots and dashes are inked on regulation telegraph recorder paper slip 3/8″ wide. These recordings resemble the conventional visually read slip, excepting the pen is heavier which results in an inked line about 1/16″ thick and the ink used is a black India drawing ink. Paper slip is mounted on 400′ 16mm, motion picture reels, each roll containing sufficient inked slip to operate the keyer at speed of 20 words per minute for one hour.

Paper slip is drawn through a guiding gate between an exciter lamp and a photo tube. The inked dots and dashes interrupt light onto photo tube, actuating a relay which keys any external tone source.

Practice tapes consist of a set of fifteen rolls prepared from master tapes furnished by Signal Corps School, Fort Monmouth, N. J. Best results will be obtained if the keyer is used in collaboration with Signal Corps School Pamphlet No. 53, Radio Operator's Manual, 1940 Edition.



RADIO TELEGRAPH SIGNAL RECORDER MODEL RRD-900

PRICE $90.00 Net

This is a device designed to make inked recordings of transmitted dots and dashes of the radio telegraph code. Amplifying circuit and copperoxide rectifier are built into the unit. Circuit diagram accompanies each instrument. Slip is drawn through recorder by tape puller such as the G-13.

T. R. McElroy

MAC

82 Brookline Ave., Boston, Mass., U. S. A.

4. Explain the purpose of using a center-tap return connection on the secondary of a transmitting tube's filament transformer.

The effective (average) potential of the filament of a filamentary type tube with respect to the grid and plate is the same as the potential at the exact center of the filament. When using alternating current to heat the filament of the tube, there will be a small A.C. voltage impressed upon the grid unless the return is made to the exact center tap.

5. If the high-voltage secondary of a plate transformer was changed from a full-wave center-tapped to a bridge rectifier connection, what would be the relative voltage and current output ratings as compared to those for the full-wave center-tapped connection?

Changing from a full-wave center-tapped to a bridge rectifier would double the power supply output voltage and cut the permissible current in half.

6. Why is it advisable to use a plate power supply for the oscillator of a transmitter separate from the final amplifier plate power supply?

To prevent final amplifier power supply voltage variations, such as might be caused during keying or modulation, from being applied to the oscillator and causing undesired frequency modulation.

7. How does a swinging choke operate to improve the voltage regulation of a plate supply filter system?

Its inductance, and hence the voltage drop across it, decreases considerably with increasing load. The high inductance at low values of current prevents soaring of the output voltage at light loads.

8. Why is full-wave rectification generally preferable to half-wave rectification in a power supply?

A full-wave rectifier supplies twice as many pulses per second to the filter, for a given supply frequency, and therefore is easier to filter and provides better voltage regulation.

9. What are the relative advantages and disadvantages of mercury-vapor and high-vacuum rectifiers of equivalent filament ratings?

A mercury-vapor rectifier has lower internal resistance than the high-vacuum rectifier, and will therefore usually supply more voltage to the filter under load (for a given transformer voltage). Because of its lower internal resistance, the mercury-vapor rectifier is more likely to be damaged from accidental overload or short circuiting. Mercury-vapor rectifiers sometimes cause "hash" to be generated because of transient oscillations set up each time the mercury vapor becomes ionized.

10. What are the principal output voltage ripple frequencies with half-wave and full-wave single-phase rectifiers, in terms of the a.c. supply frequency?

With a half-wave rectifier the principle output ripple frequency is equal to the supply-voltage frequency; with a full-wave rectifier it is equal to twice the supply-voltage frequency.

11. What is the principal reason for using a filter in a plate power supply system?

The filter in a power supply is used to smooth out the irregularities or "ripple" in the rectified alternating current, thus delivering a pure, unvarying voltage to the load circuit.

12. What would be a suitable type and the approximate capacitance of the filter condensers in a typical 1000-volt transmitter plate supply system?

With a choke-input, two-section filter, the first condenser should be at least a 2-μfd. unit and the second 4-μfd., if the power supply is to be used on a plate-modulated stage. With a radiotelegraph transmitter, the second condenser could be reduced to 2-μfd. The condensers preferably should be of the oil-filled paper type and rated at 1500 volts d.c. working voltage.

13. What would be the visible operating results of a short-circuited filter condenser in a plate power supply with an unfused primary circuit?

The first result would be a much more intense glow in mercury-vapor rectifiers or extremely high plate dissipation (as indicated by the plate becoming red hot) with high-vacuum type rectifier tubes.

14. Why should a fuse be used in the transformer primary circuit of a power supply system?

To prevent the damaging of power supply components in case of a short circuit in the power supply and to guard against the possibility of fire by preventing the power-supply components from becoming overheated in case of such a short circuit.

15. Why is a bleeder resistor connected across the output circuit of a high-voltage power supply system?

To improve the voltage regulation by keeping a small load on the power supply at all times when it is turned on and to provide a load which will discharge the filter condensers in cases where the power supply is turned off when no external load is connected.

16. What would happen if the primary of a 60-cycle power supply was connected to mains carrying continuous direct current?

The fuse would probably blow, if the primary were fused. Otherwise the primary or primaries of the transformers in the power supply would soon be burned out.

17. What is the principal advantage of a screen-grid type R.F. amplifier tube over a triode of equal output rating?

The screen-grid type tube required no neutralization when used in a properly designed r.f. amplifier stage.

18. What tube rating indicates the maxi-

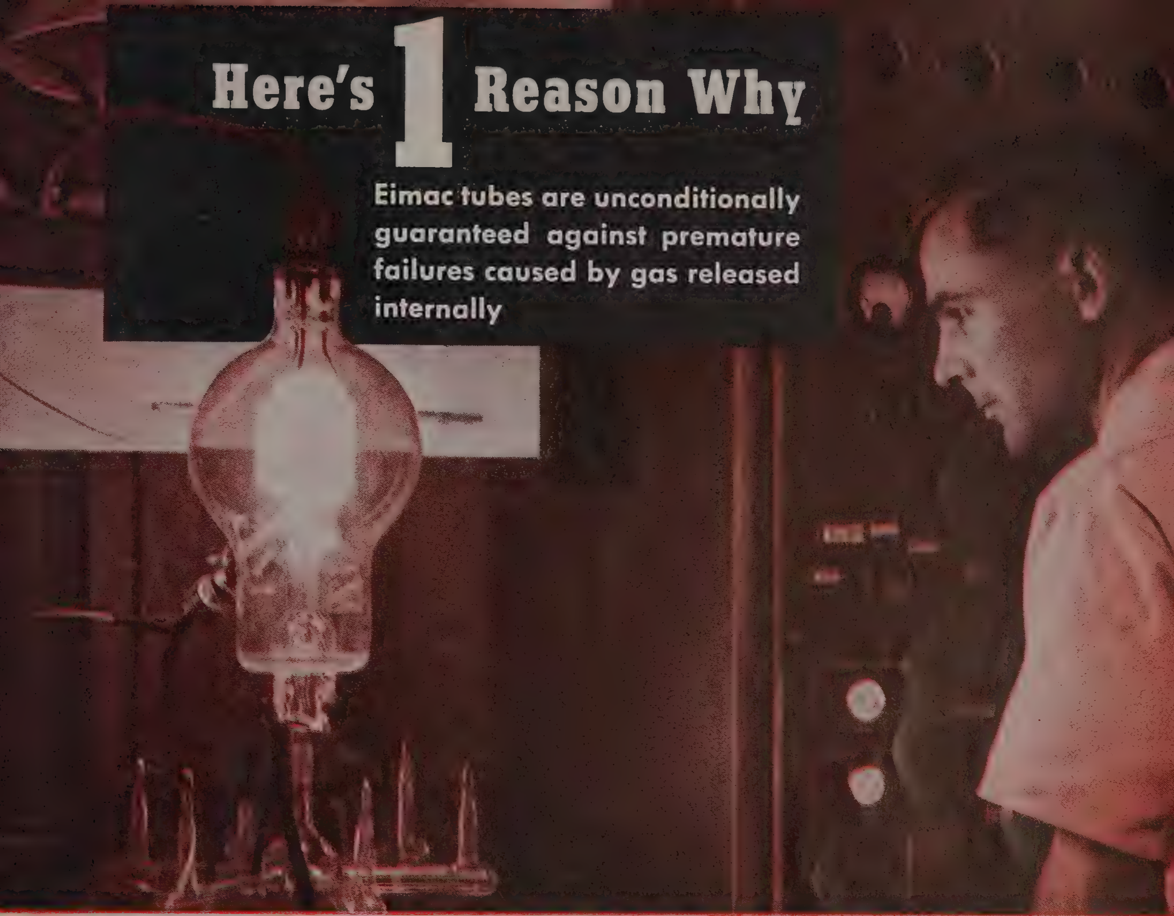
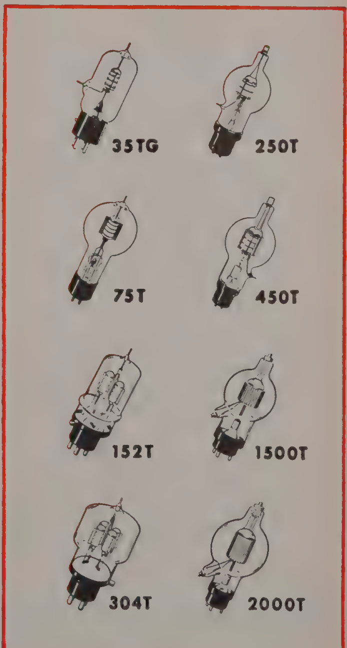


Plate dissipation is run up to ten times normal during the exclusive pumping process

Tube elements are fabricated from Tantalum which has the lowest original gas content of any known metal. This relatively small original gas content is then completely removed by the patented Eimac pumping process which you see illustrated in part above. During this process the plate dissipation is run up to ten times the normal ratings of the tube . . . and held there for hours . . . infinitely hotter than they will ever become in operation even under extreme circumstances. All Eimac tubes undergo this severe gruelling and they must pass 100% before they come to you.

Long filament life, uniformity of characteristics, outstanding performance and complete freedom from failures caused by gas released internally are basic reasons why you should adopt Eimac tubes for your transmitter. These are the same reasons why most all the world's leading engineers choose Eimac . . . and why practically every major airline and many of the most vital commercial and government transmitters throughout the world have Eimac tubes in the key sockets.



More than 20 tube types. Get information NOW. Data sheets will be sent on request.

Follow the leaders to Eimac TUBES

EITEL-McCULLOUGH, INC. San Bruno, California

mum safe heat radiation capability of the anode of a vacuum tube?

The plate dissipation rating.

19. In the classification of tubes according to the number of elements, how many grids has each of the following types:
(a) diode
(b) triode
(c) tetrode
(d) pentode
(e) heptode

(a) none
(b) one
(c) two
(d) three
(e) five

20. Describe the adjustment procedure for proper neutralization in a radio-frequency power amplifier using an r.f. indicator coupled to the plate tank circuit.

First the plate voltage lead is disconnected. With excitation applied to the stage to be neutralized and the plate circuit tuned to resonance as shown by the indicator being used, the neutralizing adjustment should be varied until the neutralizing indicator shows zero output in the amplifier plate circuit. The plate and grid circuits should be kept tuned to resonance while the neutralizing adjustment is varied.

21. Why is it necessary to neutralize a triode radio-frequency power amplifier operating with input and output circuits tuned to the same frequency?

To prevent feedback of r.f. energy from the plate to grid through the plate-to-grid capacity of the tube. Feedback may cause the tube to self-oscillate.

22. What undesirable effects may result from operation of an unneutralized triode r.f. amplifier in a transmitter?

The stage may oscillate and thus cause spurious interfering signals to be generated. If the stage is modulated the spurious signals may change frequency and strength during modulation, thus causing interference over a wide band of frequencies. Improper neutralization of a modulated stage may also cause the signal to be badly distorted even though the stage does not self-oscillate at any time.

23. What undesirable effects result from frequency modulation of an amplitude-modulated carrier wave?

Spurious signals which occupy a wide band of frequencies and cause unnecessary interference may be transmitted.

24. What operating conditions would be favorable for harmonic generation in a radio-frequency doubler or frequency multiplying amplifier?

High bias, high excitation, and a sharply peaked exciting waveform are conducive to the generation of harmonics. A single-ended stage is a better generator of even harmonics than a push-pull stage. Likewise a high mu tube is a better frequency multiplier than a low mu tube, and a tetrode or pentode a better harmonic generator than a triode.

25. Where is link coupling applicable in an oscillator-amplifier type transmitter?

Between the plate tank of the oscillator and the grid tank of the amplifier (or between any two stages). Between the amplifier plate tank and the antenna tuning or matching tank.

26. What is the purpose of a Faraday (electrostatic) shield between the output circuit of an r.f. power amplifier and antenna coupling system?

To minimize electrostatic coupling to the antenna and thus prevent harmonics from reaching the antenna by this means.

27. What are the output circuit conditions for obtaining optimum power output from a radio-frequency amplifier?

The output tank circuit should have sufficient "Q" that maximum output occurs at the point of minimum plate current. The output tank should be operated at exact resonance. The amplifier should work into the proper value of non-reactive load.

28. In which stage of a transmitter is an amplifier of high harmonic output least desirable?

In the stage that feeds the antenna.

29. What are the relative plate current indications for resonance and off-resonance running of the plate tank circuit of a radio-frequency power amplifier?

When the stage is properly designed and loaded, or is running unloaded, resonance occurs at the point of minimum plate current. Off resonance the current is quite high.

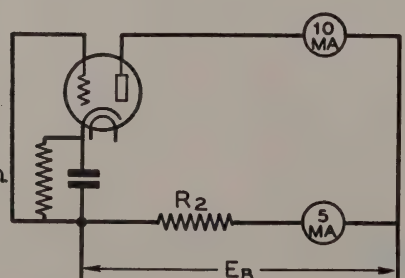
30. What are the advantages of a push-pull r.f. power amplifier output stage as compared to a single-ended stage of the same power?

The push-pull stage is easier to neutralize at high frequencies, requires less "Q" or tuning capacity in the plate tank circuit, and has much lower even harmonic output (2d, 4th, etc.).

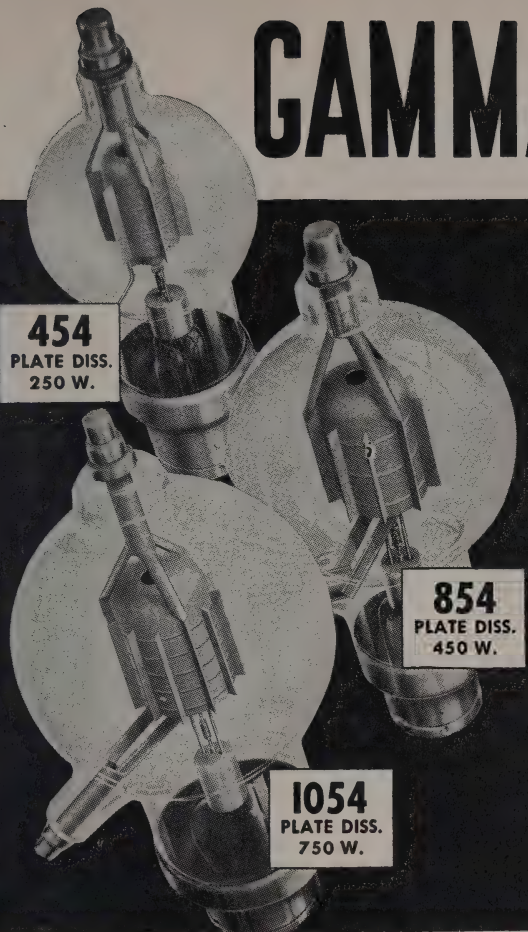
31. In the circuit diagram (left) what is the value of the bias voltage? What is the value of the bleeder resistance, R_2?

Bias voltage equals 3 volts (10 ma. or .01 amp. through 300 ohms). R_2 is equal to the plate voltage E divided by .005 (amp).

Q31



GAMMATRON TUBES



Developed in the Gammatron plant the new types 454, 854 and 1054 have copper to glass seals and special plate, grid and filament design to give a new high in UHF performance. Other features: ability to stand high plate voltages, complete protection against failure through overloads and extra long life.

Fourteen years of pioneering and experience in tantalum tubes are built into this complete line covering a power range of 50 to 5000 watts. Also available are variations of these types and high vacuum tantalum rectifiers.

The Gammatron engineers responsible for these developments will be glad to help with your special problems.

HEINTZ AND KAUFMAN LTD.
SOUTH SAN FRANCISCO CALIFORNIA U.S.A.

TYPE NO.	24	54	152L	158	254	257*	304L	354C	354E	454L	454H	654	854L	854H	1054L	1554	2054A	3054
MAX. POWER OUTPUT: Class 'C' R.F.	90	250	610	200	500	230	1220	615	615	900	900	1400	1800	1820	3000	3600	2000	5300
PLATE DISSIPATION: Watts	25	50	150	50	100	75	300	150	150	250	250	300	450	450	750	1000	1200	1500
AVERAGE AMPLIFICATION FACTOR	25	27	10	25	25		10	14	35	14	30	22	14	30	13.5	14.5	10	20
MAX. RATINGS: Plate Volts	2000	3000	3000	2000	4000	4000	3000	4000	4000	5000	5000	4000	6000	6000	6000	5000	3000	5000
Plate M.A.	75	150	500	200	225	150	1000	300	300	375	375	600	600	600	1000	1000	800	2000
Grid M.A.	25	30	75	40	40	25	150	60	70	60	85	100	80	110	125	250	200	500
MAX. FREQUENCY, Mc.: Power Amplifier	200	200	175	100	175	150	175	50	50	150	150	50	125	125	100	30	20	30
INTERELECTRODE CAP: C g—p u.u.f.	1.7	1.8	5	4.6	3.6	0.04	9	3.8	3.8	3.4	3.4	5.5	5	4	5	11	18	15
C g—f u.u.f.	2.5	2.1	7	4.7	3.3	13.8 In.	12	4.5	4.5	4.6	4.6	6.2	6	8	8	15.5	15	25
C p—f u.u.f.	0.4	0.5	0.4	1.0	1.0	6.7 Out.	0.8	1.1	1.1	1.4	1.4	1.5	0.5	0.5	0.8	1.2	7	2.5
FILAMENT: Volts	6.3	5.0	5-10	12.6	5.0	5.0	5-10	5	5	5	5	7.5	7.5	7.5	7.5	11	10	14
Amperes	3	5	13-6.5	2.5	7.5	7.5	13-26	10	10	11	11	15	12	12	21	17.5	22	45
PHYSICAL: Length, Inches	4¼	5 7/16	7¾	4¾	7	6¾	7¾	9	9	10	10	10⅜	12½	12½	16½	18	21¼	30¾
Diameter, Inches	1⅜	2	2½	2	2⅝	2⅝	3½	3⅜	3⅜	3¾	3¾	3¾	5	5	7	6	6	9
Weight, Oz.	1½	2½	8	4	6½	6	9	6½	6½	7	7	14	14	14	42	56	66	200
Base	Small UX	Std. UX	Johnson #213	Std. UX	Std. 50 Watt	Giant 7 Pin	Johnson #213	Std. 50 Watt	Std. 50 Watt	Std. 50 Watt	Std. 50 Watt	Std. 50 Watt	Std. 50 Watt	Std. 50 Watt	Johnson #214	HK 255	W.E. Co.	HK 255
NET PRICE	4.75	8.00	30.00	18.50	13.50	27.50	65.00	24.50	24.50	27.50	27.50	75.00	75.00	75.00	175.00	225.00	300.00	395.00

*Beam Pentode.

WRITE FOR FULL DATA ON ALL GAMMATRONS

32. A certain 1750-kc. Y-cut quartz crystal has a positive temperature coefficient of 125 cycles per degree Centigrade and is started in operation at 40 degrees Centigrade. If the temperature-frequency characteristic is linear, what will the oscillation frequency be at a temperature of 60 degrees Centigrade?

1752.5 kc.

Note: A positive temperature coefficient means that the crystal drifts higher in frequency with an increase in temperature.

33. A 2000-kc. low-drift crystal having a negative temperature coefficient of 5 cycles per megacycle per degree Centigrade is started in operation at 40 degrees Centigrade. If the temperature-frequency characteristic is linear, what will the oscillation frequency be at a temperature of 60 degrees Centigrade?

1999.8 kc.

34. A low-drift crystal for the 3500-4000 kc. amateur band is guaranteed by a manufacturer to be calibrated to within 0.04% of its specified frequency. Desiring to operate as close to the lower band limit of 3500 kc. as safely as possible, for what whole-number kilocycle frequency should you order your crystal, allowing 1 kc. additional for variation from temperature and circuit constants?

3503 kc.

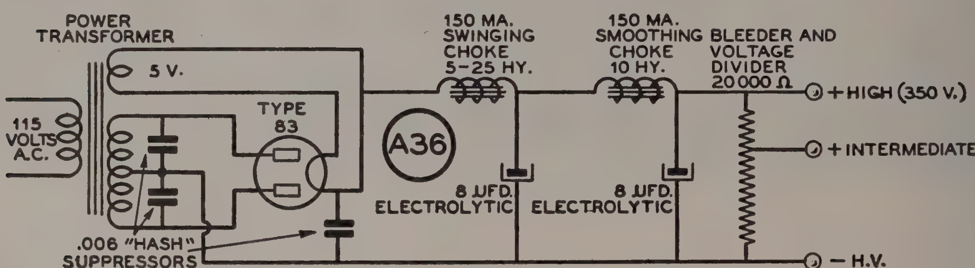
35. For what frequency should you order your crystal for operation as close as safely possible to the upper band limit of 4000 kc., with the same calibration accuracy and allowance given in Question 34?

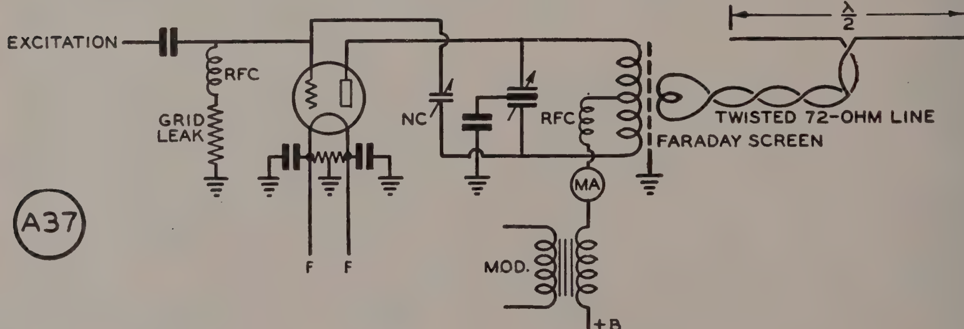
3997 kc.

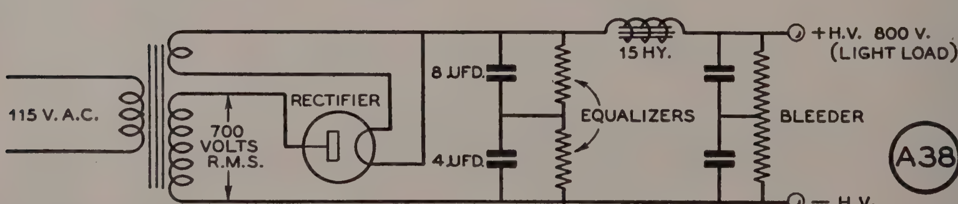
36. Draw a schematic diagram of a full-wave single-phase power supply using a center-tapped high-voltage secondary with a filter circuit for best regulation, showing a bleeder resistor providing two different output voltages and a method of suppressing "hash" interference from the mercury-vapor rectifier tubes. Give the names of the component parts and approximate values of filter components suitable for either amateur radiotelephone or radiotelegraph operation.

See diagram A36.

37. Draw a simple schematic diagram of a plate-neutralized final r.f. stage using a triode tube coupled to a Hertzian antenna, showing the antenna system and a Faraday screen to







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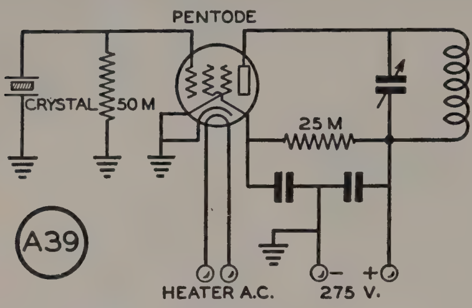


reduce harmonic radiation.

See diagram A37.

38. Draw a simple schematic diagram of a half-wave rectifier with a filter which will furnish pure d.c. at highest voltage output, showing filter condensers of unequal capacitance connected in series with provision for equalizing the d.c. drop across the different condensers.

See diagram A38.

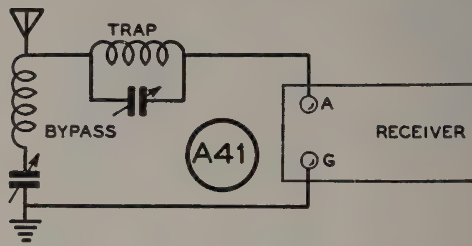
39. Draw a simple schematic diagram of a piezo-electric crystal-controlled oscillator using a pentode vacuum tube, indicating polarity of electrode supply voltages where externally connected.

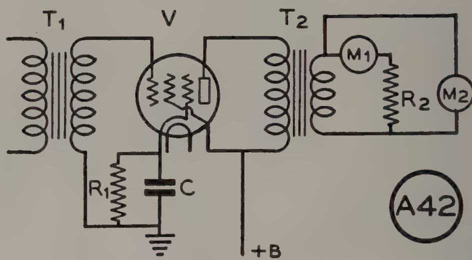
See diagram A39.

40. Draw a simple schematic diagram of two r.f. amplifier stages using triode tubes, showing the neutralizing circuits, link coupling between stages and between output and antenna system, and a keying connection in the negative high-voltage lead including a key-click filter.

See diagram A40.

41. Draw a schematic diagram of a filter for reducing amateur interference to broad-





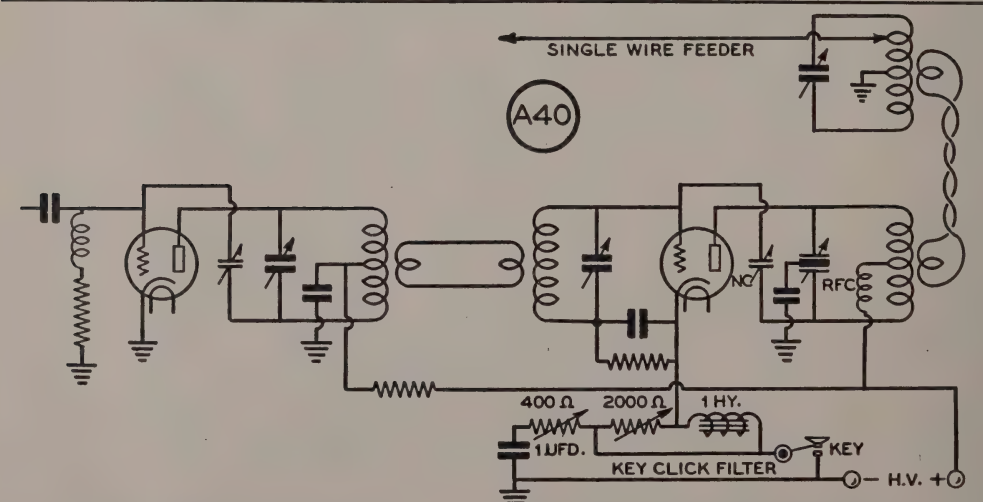
cast reception consisting of a series-tuned circuit connected in shunt with the b.c. receiver input to bypass the interfering signal and a parallel-tuned (trap) circuit in series with the receiver input to reject the interfering signal.

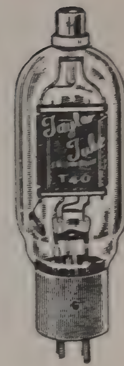
See diagram A41.

42. Draw a schematic diagram of a pentode audio power amplifier stage with an output coupling transformer and load resistor, showing suitable instruments connected in the secondary for measurement of the audio-frequency voltage and current, and naming each component part.

See diagram A42.

43. What is the principal purpose of using door interlock switches on a transmitter?





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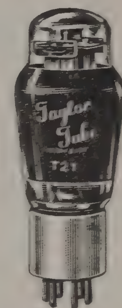
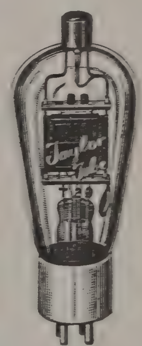


TRIODES
Class C Service

Type	Fil. Volt	Fil. Amp.	Amp. Factor	G-P mmf.	Pl. Volt	Pl. Ma.	Grid Ma	Pl. Diss. Watts	List Price
T-20	7.5	1.75	20	5.0	750	85	25	20	$ 2.25
TZ-20	7.5	1.75	62	5.0	750	85	30	20	2.25
T-40	7.5	2.5	25	4.8	1500	150	40	40	3.50
TZ-40	7.5	2.5	62	5.0	1500	150	45	40	3.50
T-55	7.5	3.0	20	3.85	1500	165	40	55	6.00
TW-75	7.5	4.15	20	1.5	2000	175	60	75	8.00
T-125	10.0	4.5	25	6.0	2500	250	70	125	13.50
TW-150	(A)	(A)	35	2.0	3000	200	60	150	15.00
T-200	10.0	5.75	17	7.9	2500	350	75	200	21.50
203A	10.0	3.25	25	14.0	1250	175	60	100	10.00
203Z	10.0	3.25	85	—	1250	175	50	65	8.00
204A	11.0	3.85	25	15.0	2500	300	80	250	60.00
211	10.0	3.25	12	14.0	1250	175	60	100	10.00
211C	10.0	3.25	12	9.0	1250	175	60	100	12.50
HD203A	10.0	4.0	25	14.0	1750	250	60	150	14.50
HD211C	10.0	4.0	12	8.0	1750	250	60	150	14.50
303C	10.0	3.25	25	8.0	1500	175	60	125	14.50
805	10.0	3.25	45	7.7	1750	200	70	125	13.50
814	10.0	4.0	12	13.5	2500	300	70	200	18.50
822	10.0	4.0	30	13.5	2500	300	70	200	18.50
845	10.0	3.25	5	14.0	1250	150	60	100	10.00
(A) Available in large or small base with 10v.-4.1a. or 5v.-8.2a. fil									
BEAM POWER TUBE									
T-21	6.3	0.9	138	1.4	400	95	5	21	1.95



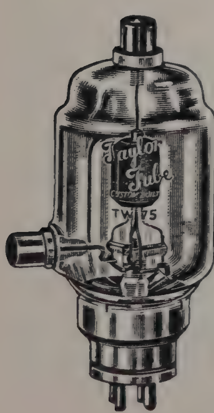
TRIODES
Class B Audio Service

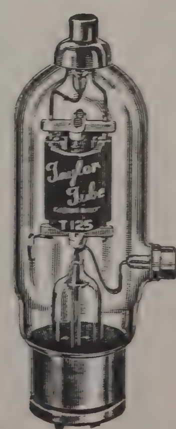
Type	Drive Power Watts	Output Watts	Pl. Volt	Pl. Ma.	Load Ohms
TZ-20	2.6	80	750	170	9000
TZ-40	6.0	250	1500	250	12000
203Z	6.75	300	1250	350	8000
805	10.0	510	1750	420	9350
822	8.0	1000	3000	450	16000

All Class B Ratings are for Two Tubes

RECTIFIERS
Half-Wave Mercury Vapor Tubes

Type	Fil. Volt	Fil. Amp.	Peak Inv. Volt	A. V. Pl. MA.	Peak Pl. Amp.	List Price
866-866 A	2.5	5.0	10,000	250	1.0	$ 1.50
866 Jr.	2.5	2.5	5,000	125	0.5	1.00
249 B	2.5	7.5	10,000	375	1.5	5.00
258 B	2.5	7.5	10,000	375	1.5	6.00
872 A	5.0	6.75	10,000	1250	5.0	10.50
875 A	5.0	10.0	15,000	1500	6.0	30.00





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Door interlock switches are used to protect the operator by opening the a.c. line voltage when the door is opened.

44. What is the usual means for protecting amateur station equipment from damage by charges of atmospheric electricity on the antenna system?

By grounding the feeders or a voltage node on the antenna through an r.f. choke or a high resistance non-inductive resistor. When not in use the antenna is grounded directly through a grounding switch.

45. What is a safe procedure for removing an unconscious person from contact with a high voltage circuit?

Turn off the voltage, or if impossible, pull him free by loose folds of clothing (if dry).

46. Using a frequency meter with a possible error of 0.75%, on what whole-number kilocycle frequency nearest the high-frequency end of the 3500-4000 kc. amateur band could a transmitter safely be set?

3970 kc.

47. Using a frequency meter with a possible error of 0.75%, on what whole-number kilocycle frequency nearest the low-frequency end of the 7000-7300 kc. amateur band could a transmitter safely be set?

7053 kc.

R. M. A. COLOR CODE

For Fixed Condensers, Unit: Micro-Microfarads

FIRST DOT		SECOND DOT		THIRD DOT	
Black	0	Black	0		
Brown	1	Brown	1	Brown	0
Red	2	Red	2	Red	00
Orange	3	Orange	3	Orange	000
Yellow	4	Yellow	4	Yellow	0000
Green	5	Green	5	Green	00000
Blue	6	Blue	6	Blue	000000
Purple	7	Purple	7	Purple	0000000
Gray	8	Gray	8	Gray	00000000
White	9	White	9	White	000000000

For Resistors, Unit: Ohms

BODY COLOR		END COLOR		DOT COLOR	
Black	0	Black	0		
Brown	1	Brown	1	Brown	0
Red	2	Red	2	Red	00
Orange	3	Orange	3	Orange	000
Yellow	4	Yellow	4	Yellow	0000
Green	5	Green	5	Green	00000
Blue	6	Blue	6	Blue	000000
Purple	7	Purple	7	Purple	0000000
Gray	8	Gray	8	Gray	00000000
White	9	White	9	White	000000000

RADIO SYMBOLS

The following symbols are commonly used in radio work and many of these symbols are used in the pages of this book:

E_F ....Filament (or heater) terminal voltage.
E_B ....Average plate voltage (d.c.).
I_B ....Average plate current (d.c.).
E_P ........A.C. component of plate voltage (effective value).
I_P ........A.C. component of plate current (effective value).
E_C ....Average grid voltage (d.c.).
I_C ....Average grid current (d.c.).
E_G ........A.C. component of grid voltage (effective value).
I_G ........A.C. component of grid current (effective value).
E_{FF} ....Filament (or heater) supply voltage.
E_{BB} ....Plate supply voltage (d.c.).
E_{CC} ....Grid supply voltage (d.c.).
M_U or μ ...Amplification factor.
r_P ....Plate resistance
S_M ....Grid-plate transconductance (also mutual conductance, gm).
R_P ....Plate load resistance.
Z_P ....Plate load impedance.
D.C. ......Direct Current.
A.C. ......Alternating Current.
RMS .....Root Mean Square.
U.P.O. ...Undistorted power output.
C_{GK} ......Grid-cathode (or filament) capacitance.
C_{PK} ......Plate-cathode (or filament) capacitance.
C_{G1P} .....Effective grid-plate capacitance in a tetrode (cathode [or filament] and screen grounded).
$C_{G1}(k+g_2)$ Direct interelectrode capacitance of grid to cathode (or filament) and screen.
$C_P(k+g)$ Direct interelectrode capacitance of plate to cathode (or filament) and screen.

α, alpha—Coefficients.
β, beta—Coefficients.
γ, gamma—Coefficients.
Δ, delta (capital)—Decrements, increments, variations.
δ, delta (lower case)—Same as capital delta
Θ, theta—Angles, phase displacement.
λ, lambda—Wavelength.
μ, mu—Amplification factor, prefix micro-.
π, pi—3.1416, circumference divided by diameter.
ϕ, phi—Angles.
τ, tau—Time constant, coefficients.
ω, omega—Resistance in ohms, 2π times frequency.

Buyer's Guide

Parts Required for Building Equipment Shown in This Book

The parts listed are some of those actually used by "Radio's" laboratory in constructing the models shown. Other parts of equal merit and equivalent electrical characteristics may usually be substituted without materially affecting the performance of the units.

CHAPTER 6
RADIO RECEIVER CONSTRUCTION

Figure 3, page 137
Two-Tube Autodyne

C_1—Hammarlund SM-15
C_2—Hammarlund SM-100
C_3, C_5—Solar type MT
C_4, C_6—Solar "Sealdtite"
R_1—Centralab 710
R_2—Centralab "Radiohm"
R_3, R_4—Centralab 514
BC—Mallory 1.25 v.
CH_1—Stancor type C-2300
Panel—Bud PS1201
Tuning dial—Bud D-103B

Figure 8, page 140
Three-Tube Simple Super

C_1, C_2—Hammarlund MC-50-S
C_3—Hammarlund MC-140-S
C_4, C_8, C_{11}, C_{14}—Cornell-Dubilier DT-4P1
C_5, C_9—Cornell-Dubilier DT-4T1
C_6, C_7, C_{12}—Cornell-Dubilier DT-4S1
C_{10}—Cornell-Dubilier DT-4D1
C_{13}—Cornell-Dubilier BR-252
C_{15}—Cornell-Dubilier EDJ-9040
R_1, R_2, R_4, R_7, R_9—Centralab 516
R_3, R_5—Centralab 514
R_6—Yaxley L
R_8, R_{10}—Ohmite "Brown Devil"
IFT—Meissner 16-8092
CH—Stancor C-2300
J—Mallory-Yaxley 705
Dial—Crowe 123M

Figure 11, page 143
Economical 5-Tube Super

C_1, C_2—Cardwell ZR-50-AS
C_3—Cardwell ZR-25-AS
C_4—Cardwell ZU-140-AS
C_5, C_7, C_8, C_9, C_{10}, C_{11}, C_{14}, C_{17}—Cornell-Dubilier DT-4P1
C_6, C_{13}—Cornell-Dubilier 5W-5T1
C_{12}—Cornell-Dubilier 5W-5T5
C_{15}—Cornell-Dubilier BR-845
C_{16}—Cornell-Dubilier BR-102-A
R_1, R_6, R_7, R_8, R_{11}, R_{13}, R_{14}—Centralab 710
R_2—Centralab 516
R_3, R_{10}—Centralab 514
R_5, R_9—Mallory-Yaxley G
R_{12}—Ohmite "Brown Devil"
IFT_1—Meissner 8091
IFT_2—Meissner 8099
Tubes—RCA

Figure 14, page 146
6K8-6J5 Converter

C_1—National ST-50
C_2—National ST-100
C_3—National ST-35
C_4, C_7, C_8, C_{10}—Aerovox 1467
C_5, C_6, C_9—Aerovox 484
R_1, R_3, R_4, R_5, R_6—Centralab 710
R_2—Centralab 514
S—Bud SW-1115
L_1, L_2 Forms—Bud 595
IFT—Meissner 16-8100
BOT—Meissner 17-8175
Chassis—Bud CB-41
Cabinet—Bud C-973
Dial—National "B"

Figure 17, page 149
High-Performance Converter

C_1, C_2, C_3—Bud 1852
C_4, C_5—Sprague 45-12
C_6—Bud LC-1682, 1683
C_7, C_8, C_9, C_{12}, C_{13}, C_{15}—Aerovox 484
C_{11}, C_{14}—Aerovox 1467
R_1—Centralab 62-113
R_2, R_3, R_4, R_5, R_6, R_7—Centralab 710

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Model 9-1081 Price on Application

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MC 28-56
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SIGNAL SPLICER
Antenna Matching Unit
SIGNAL CALIBRATOR
Frequency Standard
TRAFFIC SCOUT
9-tube Receiver
TRAFFIC MASTER
14-tube Receiver
CRYSTAL OVEN
Temperature Control
CRYSTAL FILTER
Complete Unit
B-F OSCILLATOR
Complete Unit
NOISE SILENCER
Adapter Unit
WAVE TRAPS
RELAYS
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INSULATORS
CONDENSERS
SWITCHES
I-F TRANSFORMERS

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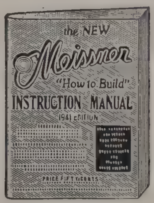
Model 9-1058 Net Price........$52.25

INSTRUCTION MANUAL



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R_8, R_9—Ohmite "Brown Devil"
S_2—Centralab 1462
T—Thordarson T-13R12
CH—Thordarson T-13C27
Coil forms—Bud 126
Tubes—RCA
Chassis—Bud 793
Shield partitions—Bud 1246, see text

Figure 19, page 151
Variable-Selectivity Crystal Filter

C_1, C_2—Aerovox 1467
C_3, C_6—Hammarlund APC
C_4, C_7, C_8—Aerovox 484
C_5—Sprague SM-31
R_1, R_2—Centralab 710
T_1, T_2—Made from Meissner 16-5740 or 16-6131, see text
X—Bliley CF-1
Cabinet—Bud 728

Figure 22, page 153
Outboard I.F. Stage

T_1, T_2—Meissner 16-6131
R_1, R_2, R_3—Centralab 710, 714
C_2, C_3, C_4—Aerovox 484

Figure 25, page 156
High-Gain Preselector

C_1, C_2—Bud MC-903
C_3—Cornell-Dubilier DT-4P1
C_4, C_5—Cornell-Dubilier DT-4S1
R_1—Centralab 514
R_2—Centralab 710
R_3—Centralab 72-113
R_4—Centralab 514
Coil sockets—Hammarlund S-5
Coil forms—Hammarlund CF-5-M and SWF-5

CHAPTER 12

EXCITERS AND LOW POWERED TRANSMITTERS

Figure 2, page 268
One-Tube Exciter

C_1, C_6—Cardwell ZR-50-AS
C_2, C_3—Aerovox 1450
C_4, C_5, C_7—Aerovox 684
R_1—Centralab 516
R_2, R_3, R_4—Ohmite "Brown Devil"
RFC—Bud 920
X—Bliley LD2
Coil forms—Hammarlund XP-53
Tube—Taylor T21

Figure 3, page 268
Power Supply for Figure 2

T—Thordarson T-13R13
CH—Thordarson T-57C53
C—Sprague LR-88
R—Ohmite "Brown Devil"

Figure 5, page 269
160-Meter 5-Watt V.F.O.

C_1—National STH-335
C_2—National SS-150
C_3—Sprague SM-31
C_4, C_6, C_9—Sprague 1FM26
C_5, C_7, C_8—Sprague TC-15
R_1—Centralab 710
R_2, R_3, R_4—Centralab 714
RFC—National R-100
Coil form—National XR-13
Dial—National type ACN

CHAPTER 13
MEDIUM AND HIGH POWER
R.F. AMPLIFIERS

Figures 2 and 3, pages 287 and 288

C_1—National TMS-50D
C_2—National TMH-75D
C_3, C_4—National NC-75
C_5, C_6—Solar MW-1235
C_7—Solar XM-25-22
R_1—Ohmite "Brown Devil"
L_1—National AR-16 series

Figure 11, page 273
Shuart 25-Watt V.F.O.

Osc. grid tank—Hammarlund ECO-160
Osc. plate tank—Hammarlund ETU-80
Buffer tank—Hammarlund 6014
C_1—Hammarlund MC-140-S
C_5—Hammarlund MC-100-S
RFC—Hammarlund CHX
L_4 coil forms—Hammarlund SWF-5
807 Shield—Hammarlund PTS
Ceramic sockets—Hammarlund type S
T—Thordarson type 75R50
CH—Thordarson 75C51
Dial—National type N
Tubes—RCA throughout

Figure 14, page 275
Cascade Frequency Multiplier

C_1, C_2—Cardwell ZR-35-AS
C_3, C_4—Cardwell ZR-25-AS
C_5 to C_{12}—Solar type MW
R_1 to R_6—Centralab 516
R_7—Ohmite "Brown Devil"
S_1—Bud SW-1005
S_2—Centralab 1405
Chassis—Bud CB-997
Panel—Bud PS-1202
Tubes—RCA throughout

Figure 17, page 277
807 Utility Unit

C_1—Hammarlund MC-100M
C_2—Hammarlund MTC-350C
C_3, C_5—Cornell-Dubilier 1-W
C_4—Cornell-Dubilier DT-4S1
C_6—Cornell-Dubilier type 4-12050
R_1—Centralab 714
R_2, R_3, R_4—Ohmite "Brown Devil"
T—Thordarson 19F81
Coils—Bud OEL
Sockets—Johnson 225
Crystal—Bliley BC3
807—General Electric
RFC—Bud CH920S

Figure 22, page 280
814 Bandswitching Exciter

R_4, R_7, R_{14}—Ohmite, "Brown Devil"
R_{13}, PC—Ohmite P-300
C_1—Cardwell EU-140-AD
C_2—Cardwell EU-100-AD
C_3—Cardwell MT-100-GS
C_4 to C_{13}—Solar MO and MW
C_{14}—Solar XM-25-22
C_{15}—Cardwell JD-50-OS
T_1—Kenyon T-351
T_2—Kenyon T-365
Coil turret—Barker & Williamson type 2-A
S_1—Centralab 1461
S_2—Centralab 1460
S_3—Heintz & Kaufman 892
S_4—Mallory-Yaxley 151-L
Chassis—Par-Metal C-4526
Panel—Par-Metal G-3606
Crystals—Bliley B5 and LD2
Tubes—RCA throughout

L_1—Socket—National XB-16
L_2—Plug—National PB-15
L_2—Socket—National XB-15
RFC—National R-154U
Tube sockets—National CIR-4
Tube caps—National SPP-9
Dials—National "O"
Chart frame—National, size "B"
Chassis—Par-Metal C-4528
Panel—Par-Metal 3604
Feed-through insulators—Johnson 50
Tubes—RCA

Figures 1 and 4, pages 285 and 289
400-Watt Amplifier

C_1—Hammarlund MCD-100-S
C_2—Hammarlund HFBD-65-E
C_3, C_4—Hammarlund N-10
C_5, C_6—Aerovox 1450
L_1—Barker & Williamson MCL series
L_2—Barker & Williamson TVL series
Tubes—Eimac

Figures 1 and 5, pages 285 and 290

C_1—Johnson 150FD20
C_2—Johnson 150DD70
C_3, C_4—Johnson 6G70
RFC—Johnson 752
Tube sockets—Johnson 211
L_1 socket—Johnson 225
L_2 supports—Johnson 67
L_2 coil jacks—Johnson type 70
L_2 coil plugs—Johnson type 71
Control handles—Johnson 204
Tubes—H & K type HK-254

Figures 1 and 6, pages 285 and 291
1-Kw. Amplifier

C_1—Bud 1576
C_2—Bud 1818
C_3, C_4—Bud 1000
C_5, C_6—Solar XM-6-24
L_1—Bud VCL series
L_2—Bud MCL series
RFC—Bud 568
Sockets—Bud 226
Tubes—Taylor

Figures 7, 8, and 9, pages 293 and 294
Single-Ended Amplifier

C_1—Cardwell MT-100-GS
C_2—Cardwell XG-50-XD
C_3—Bud 1519
C_4, C_5, C_6—Aerovox 1450
C_7—Aerovox 1457
L_1—Barker & Williamson BL series
L_2—Barker & Williamson HDVL series
R—Ohmite 50 Watt
RFC—Hammarlund CH—500
T—Thordarson T-19F96
Socket—Johnson 213

CHAPTER 14
SPEECH AND MODULATION EQUIPMENT

Figure 3, page 297
25-Watt Modulator

Tubular condensers—Aerovox 484
C_2, C_8—Aerovox PRS450 12 μfd.
C_3—Aerovox 1467 mica
C_5, C_7—Aerovox PB-10-10 25 volt
C_9—Aerovox 600-LU 4 μfd.
C_{10}—Aerovox GL-475 8 μfd.
Carbon resistors—Centralab 1 watt
Wirewound resistors—Ohmite "Brown Devil"

R_5—Mallory-Yaxley M control
T_4—Stancor A-4721
T_2—Stancor A-3892
T_3—Stancor P-3005
CH—Stancor C-1001
Bias cell—Mallory-Yaxley
Tubes—RCA throughout

Figure 5, page 299
60-Watt T-21 Modulator

C_1, C_4—Solar S-0240
C_2—Solar S-0215
C_3, C_7—Solar S-0263
C_5, C_{13}, C_{14}, C_{15}—Solar LG5 8-8
C_{11}, C_{12}—Solar M116
C_8—Solar M010
R_1—Centralab 72-116
All ½-watt resistors—Centralab 710
All 1-watt resistors—Centralab 714
R_{15}, R_{16}, R_{21}—Centralab 516
R_{17}, R_{18}, R_{19}, R_{20}—Ohmite "Brown Devil"
BC—Mallory-Yaxley Bias Cell
T_1—Thordarson T-84D59
T_2—Thordarson T-11M75
T_3—Thordarson T-79F84
T_4—Thordarson T-84P60
CH_1—Thordarson T-75C49
CH_2—Thordarson T-75C51
CH_3—Thordarson T-68C07
Tubes—RCA 6J5, 6L7, 83, 45. Taylor T-21

Figure 8, page 301
6-Watt 6L6 Grid Modulator

C_1, C_4, C_7—Cornell-Dubilier EDJ-3100
C_2, C_8, C_9—Cornell-Dubilier BR-845
C_3—Cornell-Dubilier DT-681
C_5—Cornell-Dubilier DT-4P1
C_6—Cornell-Dubilier SM-6S5
1-watt resistors—Centralab 714
½-watt resistors—Centralab 710
R_{12}, R_{13}, R_{14}—Ohmite "Brown Devil"
R_6—Centralab 72-105 potentiometer
T_1—Stancor A-4406
T_2—Stancor P-3005
CH—Stancor C-1421
Feed-thru insulators—Bud I-436
Chassis—Bud CB-1194
Pilot light—Mallory-Yaxley 310R
Tubes—RCA throughout

Figure 10, page 303
Push-Pull 2A3 Amplifier-Driver

C_1, C_3, C_5—Aerovox 484
C_2, C_6—Aerovox PBS-25 10
C_4, C_7—Aerovox PBS-5 8-8
C_8, C_9—Aerovox WG-5 8
R_1, R_2, R_3, R_4—Centralab 710
R_5—Centralab 72-105
R_6—Centralab 710
R_7, R_9—Ohmite "Brown Devil"
Input trans.—Stancor A-72-C
T—Stancor P-4049
CH—Stancor C-1421
Tubes—RCA throughout

Figure 14, page 306
Dual-channel 6A3, P.A. or Speech Amplifier

C_1—Cornell-Dubilier BR-102A
C_2—Cornell-Dubilier DT-4P1
C_3—Cornell-Dubilier BR-845
C_4—Cornell-Dubilier 3L-5D5
C_5—Cornell-Dubilier BR-252A

C_6, C_7—Cornell-Dubilier DT-6S1
C_8—Cornell-Dubilier BR-4015
C_9—Cornell-Dubilier TL-6040
C_{10}—Cornell-Dubilier BR-845
R_1, R_4, R_5, R, R_6, R_8—Centralab 710
R_9, R_{10}, R_{11}, R_{12}, R_{13}, R_{14}, R_{16}—Centralab 710
R_2—Centralab 72-119
R_3—Centralab 72-105
R_7—Centralab 72-105
R_{15}—Ohmite "Brown Devil"
T_1—Thordarson T-13R15
T_2—Thordarson T-67S54
CH—Thordarson T-57T54
S_2—Centralab 1461

Figure 16, page 307
Class B 809 Modulator

T_1—Stancor A-4762
T_2—Stancor A-3894
T_3—Stancor P-3064
Tubes—General Electric

Figure 20, page 311
TZ-40 Modulator

T_1—Thordarson 81D42
T_2—Thordarson 15D79
T_3—Thordarson 11M77
T_4—Thordarson 70R62
T_5—Thordarson 16F13
CH_1—Thordarson 74C29
CH_2—Thordarson 13C28
All tubular condensers—Cornell-Dubilier DT
All filter condensers—Cornell-Dubilier EDJ
Tubes—RCA. TZ-40's Taylor

Figure 22, page 313
203Z Modulator

All tubular condensers—Cornell-Dubilier DT
All resistors—I.R.C. BT-½ and BT-1
R_7—Mallory-Yaxley O control
R_{14}—Mallory-Yaxley Y50MP
T_1—Thordarson T-57A41
T_2—Thordarson T-75D10
T_3—Thordarson T-11M77
T_4—Thordarson T-19F96
203Z—Taylor, Rest—RCA
Sockets for 203Z—Johnson 211

CHAPTER 15
POWER SUPPLIES

Figure 14, page 324
Voltage Regulated Power Supply

C_1—Solar M-508
C_2—Solar M-408
R_1—Centralab 710
R_2, R_3, R_4, R_6—Centralab 714
R_5—Centralab 72-115
T—Thordarson T-13R14
CH—Thordarson T-57C53
Tubes—RCA

Figure 16, page 325
Voltage Regulated Bias Pack

C_1, C_2, C_3, C_4—Solar DAA-704
R_1—Centralab 72-103
R_2, R_3, R_4, R_5, R_6, R_7—Centralab 710
R_8, R_9—Centralab 714
T—Thordarson T-13R19
S_2—Centralab 1401

Figure 28, page 330
350-Volt Power Supply

T—Thordarson T-13R14
CH—Thordarson T68C07

C—Cornell-Dubilier EH-9808
R—Ohmite "Brown Devil"
83—Hytron

Figure 31, page 331
500-Volt Power Supply

Transformers—Stancor P-3699 and P-5009
Chokes—Stancor C-1401 and C-1411
Condensers—Cornell-Dubilier TLA-6040
83—Hytron

Figure 32, page 332
Rack Mounted Supply

Transformers—Kenyon T-line
Chokes—Kenyon T-line
Bleeders—Ohmite "Dividohm"
866's—General Electric
Condensers—Cornell-Dubilier TJU-200-20 and PE-CH-4008

Figure 35, page 335
1500-Volt Power Supply

T_1—Kenyon T-672
T_2—Kenyon T-360
CH_1—Kenyon T-512
CH_2—Kenyon T-176
C_1, C_2—General Electric 26F194
866A/866's—General Electric
R—Ohmite 0924

Figure 36, page 337
Modulator and Power Supply

Transformers—Kenyon T-line
Chokes—Kenyon T-line
TZ-40's—Taylor
866's—Taylor
Condensers—Aerovox

Figure 37, page 338
Dual Power Supply

Transformers and Chokes—Thordarson "19" type
Tubes—GL-866 and RCA 83
Condensers—Mallory

Figure 38, page 338
Compact Power Supply

Transformers and Chokes—Thordarson "CHT" type
Condensers—Aerovox
Bleeders—Ohmite "Dividohm"
Tubes—RCA-866

CHAPTER 16
TRANSMITTER CONSTRUCTION

Figure 3, page 342
Exciter-Transmitter R.F. Section

C_1, C_2—Hammarlund MC-325-M
C_3—Hammarlund MTCD-25-C
C_4, C_5, C_6—Sprague TC-11
C_7—Sprague 1FM-21
C_8, C_9, C_{12}, C_{13}—Sprague 1FM-24
C_{10}—Sprague 1FM-35
R_1, R_2, R_4, R_6—Ohmite "Brown Devil"
R_3, R_5, R_7—Centralab 516
RFC_1, RFC_2, RFC_3—Hammarlund CHX
S_1—Centralab 1462
X—Bliley LD2
6L6's—RCA
HY69—Hytron
Bias battery—Burgess B30
Coil turret—Bud XCS-1
Chassis—Bud 772
Panel—Bud 1254
Panel brackets—Bud 460
Cabinet—Bud CR-1743

Figure 6, page 344
Speech Amplifier-Modulator

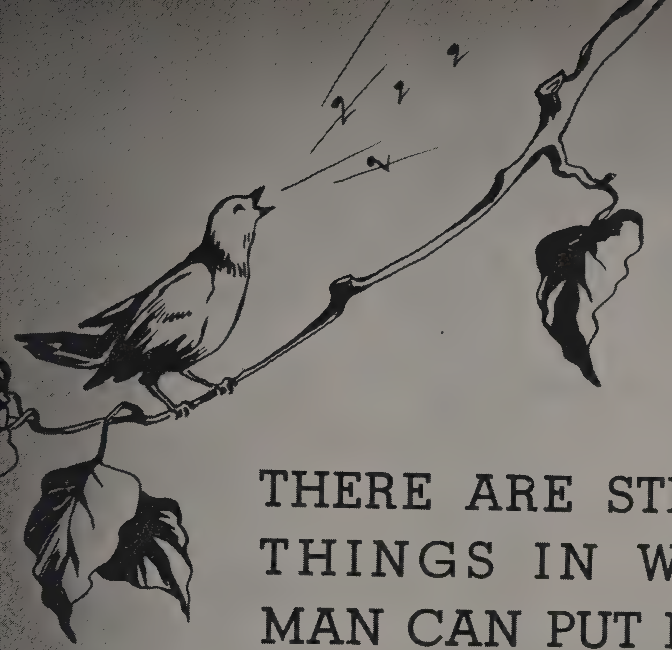
C_1—Sprague 2FM-31
C_2, C_7—Sprague TA-10
C_3—Sprague TC-11
C_4—Sprague TC-2
C_5, C_8, C_{11}—Sprague UT-8
C_6, C_9—Sprague TC-15
C_{10}—Sprague TA-510
R_1 to R_6, inclusive—Centralab 710
R_7, R_8, R_9—Centralab 714
R_{10}—Centralab 72-121
R_{11}, R_{12}—Centralab 516
R_{13}—Ohmite "Brown Devil"
T_1—Kenyon T-254
T_2—Kenyon T-493
T_2—Driver transformer—Kenyon T-271
Tubes—RCA throughout
Chassis—Bud CB-1762
Panel—Bud 1254

Figure 7, page 345
Power Supply

T_1—Kenyon T-655 (Use "low" pri. tap)
T_2—Kenyon T-367
CH_1, CH_3—Kenyon T-153
CH_2—Kenyon T-152
C_1, C_2—Sprague PC-46
C_3, C_4—Sprague SC-8
C_5—Sprague UT-161
R_1, R_3—Centralab 516
R_2, R_4—Ohmite "Brown Devil"
Tubes—Hytron

Figure 9, page 346
R. F. Amplifier and Modulator

C_1—Bud 912



C_2—Bud BC-1629
C_3, C_4, C_5—Aerovox 1450
C_6, C_7—Bud MC-567
C_8—Aerovox 1457
C_9, C_{10}—Aerovox 1509
R_1, R_2—Ohmite "Brown Devil"
R_3, R_4, R_5—Centralab 714
RFC—Bud 569
L_1—Bud OLS series
L_2—Bud VLS series
L_2—Coupling and jack assembly—Bud AM-1352
T_1—Stancor A-3894
T_2—Stancor P-6309
T_3—Stancor P-3060
T_4—Stancor P-6152
RY—Staco T-10E
S_1—Centralab 2542
S_3—Bud 1270
PC—Ohmite P-300
811's—General Electric
812's—General Electric
R.F. Chassis—Bud 643
R.F. Panel—Bud 1257
Modulator and Power Supply
Chassis—Bud CB-1762
Panel—Bud 1256
Cabinet—Bud CR-1744
812, 866 Sockets—Johnson 210
Feed-through insulators—Johnson 44
866's—General Electric

Figure 16, page 351
150-Watt C.W. Transmitter

C_1—Bud MC-905
C_2—Bud JC-1534
C_3—Bud MC-907

C_4, C_6, C_9, C_{10}, C_{12}. C_{13}—Aerovox 1467
C_5—Aerovox 484
C_7, C_8—Aerovox 1450
C_{11}—Aerovox 1446
C_{14}, C_{15}—Aerovox PBS450
C_{16}—General Electric 23F71
C_N—Bud MC-567
R_1, R_2, R_3, R_4—Centralab 514
R_5, R_6, R_7, R_8—Ohmite "Brown Devil"
R_9, R_{10}—Centralab 516
R_{11}, R_{12}—Ohmite 0585
R_{13}—Centralab 514
T_1—Inca J-13
T_2—Inca B-46
T_3—Inca J-31
CH_1—Inca D-40
CH_2—Inca D-2
S_1—Bud SW-1270
S_2—Bud SW-1115
S_3—Bud SW-1119
S_4—Mallory 151-L
S_5—Centralab 1405
L_2, L_3 Forms—Bud CF-126
L_4, L_5—Bud RCL series
L_6 Forms—Bud CF-595
X—Bliley LD2 and B5
RFC—Bud CH-920S
Cabinet—Bud CR-1742
Panel—Bud PS-616
Chassis—Bud CB-662
Feed-through insulators—Johnson 42
5Z3's—RCA
807 and 812—General Electric

Figure 24, page 359
250-Watt 'Phone-C.W. Transmitter

C_A—Cardwell XG-50-XD
C_1, C_2—Cardwell ZR-50-AS
C_3—Cardwell ZR-35-AS
C_4—Cardwell ZT-15-AS
C_5, C_6—Cardwell MT-70-GS
C_7, C_8, C_9, C_{10}—Aerovox 1467
C_{11}, C_{12}, C_{13}—Aerovox 1450
C_{14} to C_{21}, inclusive—Aerovox 1467
C_{22}—Aerovox 1457
C_{23}—Cornell-Dubilier BR-255
C_{24}, C_{25}—General Electric Pyranol
C_{26}—Aerovox 1467
C_{27}, C_{37}—Solar S-0256
C_{28}, C_{30}, C_{33}—Solar MW-1210
C_{29}—Solar S-0223
C_{31}, C_{34}—Solar S-0238
C_{32}—Solar DT-883
C_{35}—Solar DT-859A
C_{36}, C_{39}—Solar DT-879
C_{38}, C_{40}—Solar D-820
C_{41}—General Electric 23F167
C_{42}—Solar XM-25-22
C_{43}, C_{44}—General Electric 23F194
R_1, R_2, R_3, R_4, R_5, R_{19}, R_{20}, R_{21}, R_{23} to R_{29}, inclusive, R_{31}, R_{32}, R_{33}, R_{34}—Centralab 710
R_6—Centralab 514
R_7 to R_{14}, inclusive, R_{16}, R_{35}—Centralab 516
R_{17}, R_{18}, R_{30}, R_{36}—Ohmite "Brown Devil"
R_{22}—Centralab 72-116
R_{37}—Ohmite 0625
T_1—Thordarson T-19F96
T_2—Thordarson T-19F76
T_3—Thordarson T-84P60
T_4—Thordarson T-19F81
T_5—Thordarson T-19F99
T_6, T_8—Thordarson T-19F90
T_7—Thordarson T-19P60
T_9—Thordarson T-17AO2
T_{10}—Thordarson T-15D77

T_{11}—Thordarson T-11M76
CH_1, CH_2—Thordarson T-75C51
CH_3—Thordarson T-15C31
CH_4—Thordarson T-16C07
CH_5—Thordarson T-19C36
CH_6—Thordarson T-19C43
RFC_1, RFC_2—Hammarlund CHX
S_1—Centralab 2543
S_2—Centralab 2505
S_3—Centralab 2544
S_4, S_5—Bud SW-1115
S_6, S_8—H & H
S_7—Bud 1270
Rack—Bud CR-1745
Panels—Bud
Chassis—Bud 773, 643, and 770
Brackets—Bud 451 and 460
813 Socket—Johnson 237
813—General Electric
866/866A's—General Electric
Crystal—Bliley LD2
RY_1—Ward-Leonard 507-507
RY_2—Ward-Leonard 507-503

Figure 34, page 367
300-Watt 'Phone, 600-Watt C.W. Transmitter

C_1—Bud MC-910
C_2, C_3—Bud MC-905
C_4—Bud MC-1865
C_5—Bud JC-1555
C_6—Bud BC-1629
C_7, C_8, C_9—Cornell-Dubilier DT-6S1
C_{10}—Cornell-Dubilier 4-13010
C_{11}, C_{12}, C_{13}—Cornell-Dubilier DT-6S1
C_{14}—Cornell-Dubilier 4-12020
C_{15}—Cornell-Dubilier 4-13010
C_{16}—Bud NC-890
C_{17}—Cornell-Dubilier 4-22020
C_{20}, C_{21}—Cornell-Dubilier 4-12050
C_{22}, C_{23}—Cornell-Dubilier EB-9080
C_{24}—Cornell-Dubilier TQ-10040
C_{25}—Cornell-Dubilier TQ-10020
C_{26}—Cornell-Dubilier 5W-5Q5
C_{27}—Cornell-Dubilier BR-102-A
C_{28}, C_{29}, C_{30}—Cornell-Dubilier DT
C_{31}—Cornell-Dubilier BR-102-A
C_{32}—Cornell-Dubilier BR-845
C_{33}, C_{34}—Cornell-Dubilier 5W-5Q5
C_{35}, C_{36}—Cornell-Dubilier DT-4P1
C_{37}—Cornell-Dubilier BR-505
C_{38}—Cornell-Dubilier EB-9080
C_{39}—Cornell-Dubilier 4-52030
C_{40}—Cornell-Dubilier TLA-6040
C_{41}—Cornell-Dubilier EB-9080
C_{42}, C_{43}—Cornell-Dubilier TQ-20020
C_{44}—Cornell-Dubilier TJU-15040
C_{45}—Cornell-Dubilier TLA-10020
C_{46}—Cornell-Dubilier MD-8P15
R_1, R_{13}, R_{14}, R_{15}, R_{16}, R_{17}—Centralab 710
R_2, R_5, R_7, R_{11}, R_{12}, R_{28}, R_{29}, R_{31}—Ohmite "Brown Devil"
R_3, R_4, R_6—Centralab 516
R_8, R_9, R_{19}, R_{20}, R_{21}, R_{27}, R_{25}, R_{33}—Centralab 714
R_{10}—Ohmite 0215
R_{18}—Centralab 72-105
R_{30}—Ohmite 0583
R_{34}—Ohmite 0925
S_1, S_2—Centralab 1401
S_3—Centralab 1462
S_4, S_5, S_6, S_7—Arrow-H&H toggles
S_8—Centralab 1462
S_9, S_{10}—Arrow-H&H toggles
L_4, L_5—Bud OLS series
L_6—Bud VCL or VLS series
Crystals—Bliley LD-2 and B-5
RFC_1, RFC_2—Bud CH-920S
RFC_3—Bud CH-568
T_1—U.T.C. S-61
T_2—U.T.C. S-59
T_3—U.T.C. S-44
T_4—U.T.C. S-72
T_5—U.T.C. S-46
T_7—U.T.C. S-9
T_8—U.T.C. VM-3 (S-22 also suitable)
T_9—U.T.C. S-61
T_{10}—U.T.C. R-4
T_{11}, T_{12}—U.T.C. S-57
T_{13}—U.T.C. S-49
T_{14} (T_{15}, T_{16} also)—U.T.C. R-1
CH_1—U.T.C. S-30
CH_2—U.T.C. S-29
CH_3—U.T.C. S-32
CH_4—U.T.C. S-31
CH_5—U.T.C. S-29
CH_6—U.T.C. S-36
CH_7—U.T.C. S-35
CH_8—U.T.C. S-34
RY_1, RY_3—Ward-Leonard 507-510
RY_2—Ward-Leonard 507-505
PC—Ohmite parasitic choke

Figure 41, page 372
400-Watt 'Phone Transmitter

All variable condensers—Bud
All mica fixed condensers—Cornell-Dubilier type 9
All paper by-pass condensers—Solar "Domino"
Electrolytic condensers—Mallory-Yaxley
Ceramic sockets—Hammarlund type S
All wirewound resistors—Ohmite

All carbon resistors—Centralab insulated type
Tubes—Heintz & Kaufman, Ltd. HK-254's or Eimac 100TH's, HK-54 or Eimac 35T, Taylor 203Z's. All others RCA
RFC—Bud type 920
RFC_1—Bud type 569
R_{13}, R_{21}—Yaxley universal type
Tuning dials—Bud type 165
Coil forms—Bud type 126
C_{33}, C_{34}, C_{35}—Mallory oil type
Transformers—Thordarson as follows:
T_1—19F83
T_2—19F85
T_3—33A91
T_4—75D10
T_5—11M77
T_6—75R50
T_7—19F96
T_8—19P59
T_9—19F90
T_{10}—19P62
T_{11}—19F90
CH_1CH_2—19C42
CH_3—19C36
CH_4—19C43
CH_5—19C36
Crystal—Bliley LD2 or B5
Other tubes—RCA

CHAPTER 18
U.H.F. RECEIVERS AND TRANSCEIVERS

Figure 2, page 388
Five-Meter Converter

C_1, C_2—Rebuilt Cardwell ER-25-AD, see text
C_3, C_9—Cornell-Dubilier 1W-5S1
C_4, C_5—Cornell-Dubilier 5W-5T5
C_6—Meissner 22-7002
C_7—Hammarlund APC-25
C_8—Hammarlund HF-15
C_{10}—Cornell-Dubilier EDJ-9080
R_1, R_5—Centralab 710
R_2, R_3, R_4—Centralab 714
Chassis and cabinets—Bud 870-A
Tubes—RCA

Figure 4, page 389
U.H.F. Superregenerator

C_1—Johnson type 15J12 (altered)
C_2—Cornell-Dubilier type 5W-5Q5
C_3—Cornell-Dubilier 1W-3D6
C_4, C_5—Cornell-Dubilier BR-102-A
C_6—Cornell-Dubilier DT-4S5
R_1, R_2, R_3—Centralab 710
R_4—Centralab 72-122
R_5—Centralab 516
T—Thordarson 13A35
Tubes—RCA
Dial—National type 0

Figure 9, page 392
112-Mc. F.M.—A.M. Superhet

C, C_3, C_4, C_{12}, C_{17}, C_{18}, C_{28}—Sprague 2FM-31
C_1—Johnson 7J12
C_2—Johnson 15J12
C_5, to C_{11}, inclusive, C_{13}, C_{14}, C_{15}, C_{19}, C_{21}, C_{26}—Sprague TC-11
C_{16}—Sprague 2FM-45
C_{20}, C_{27}—Sprague TA-10
C_{22}—Sprague TC-1
C_{23}—Sprague TX45-35
C_{24}, C_{29}—Sprague 2FM-35
C_{25}—Sprague UT-8
R_1, R_3, R_4, R_5, R_6, R_8, R_9, R_{10}, R_{12}, R_{13}, R_{14}, R_{15}, R_{18}, R_{19}, R_{20}, R_{21}, R_{22}, R_{26}, R_{30}—Centralab 710
R_2, R_7, R_{11}, R_{16}, R_{23}, R_{24}—Centralab 714
R_{17}—Mallory-Yaxley G
R_{27}, R_{29}—Ohmite "Brown Devil"
R_{28}—Mallory-Yaxley N
T_1, T_2, T_3, T_4—Meissner 16-4261
S_1—Mallory-Yaxley 8
RFC—Hammarlund CHX
Tubes—RCA throughout

Figure 19, page 398
112-Mc. Receiver

C_1—Johnson 7J12
R_1, R_3—Centralab 710, 714
R_2, R_4—Centralab Midget Radiohm
R_5—Sprague 5-K "Koolohm"
BC—Mallory bias cell
T_1—Thordarson T-13A35
Cabinet—Bud CU-728
RFC—Bud CH-925

Figure 22, page 400
224-Mc. Receiver

C_1—Modified Cardwell "Trim-Air"
C_2—Aerovox 1468
C_3, C_4—Aerovox type 84
C_5, C_6—Aerovox type PRS
C_7—Aerovox 1467
R_1, R_2, R_3—Centralab 710
R_4—Centralab 714
R_5—Centralab 72-122
R_6—Ohmite "Brown Devil"
T_1—Thordarson T-13A35
HY615 and 6J5GT—Hytron
6F6—RCA

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Figure 24, page 401
112-Mc. Mobile Transceiver

C_1—Cardwell ZV-5-TS with unsplit stator
C_2—Sprague 2FM-31
C_3, C_4—Sprague TC
C_5—Sprague TA-10
C_6—Sprague UT-8
R_1—Centralab 516
R_2, R_4, R_8—Centralab 714
R_3—Centralab 710
R_5, R_6—Mallory-Yaxley type L
RFC—Ohmite Z-1
S_1—Centralab type 1450
T_1—Thordarson T-72A59
T_2—Thordarson T-13S38
Power supply—Mallory Vibrapack

Figure 28, page 403
112-Mc. Transceiver

C_1—Bud LC-1640 with one plate removed
C_2—Solar type MW
C_3, C_4, C_5—Solar "Sealdtite"
C_6—Bud MT-833
R_1, R_3, R_4—Centralab 710
R_2—Centralab 72-104
CH—Thordarson 13C26
T—Thordarson 72A59
S_1—Bud SW-1119
S_2—Centralab 1450
RFC—Ohmite Z-O
Cabinet—Bud 993
Panel—Bud PS1201
Dial plate—Bud DP1176
Name plates—Bud N1150, DP1180, DP1182
Knobs—Bud K581, K579, K182

Shaft coupling—Bud FC795
Bearing—Bud PB532
Feed-through insulators—Bud I1911
Batteries—(A) Burgess 2F; (B) Burgess M-30; (C) Burgess 5360

CHAPTER 19
U.H.F. TRANSMITTERS

Figure 2, page 406
HY75 112-Mc. Oscillator

C_1—Johnson 15J12
C_2—Solar type MO
RFC—Bud CH-925
Tube—Hytron
R_1—Centralab 516

Figure 4, page 407
10 W. 112-Mc. Transmitter

T_1—Thordarson type T-86A02
T_2—Thordarson T-17M59
T_3—Thordarson T-70R62
CH—Thordarson T-57C53
C_1—Cornell-Dubilier type EDJ2250
C_2, C_3—Cornell-Dubilier EH 9808 (one)
R_1, R_2, R_3—Centralab 514
R_4, R_5—Centralab 516
Tubes—RCA

Figure 6, page 408
75-T Oscillator

C_1—Bud MC-902
C_2—Aerovox 1457
C_3—Aerovox 1467
C_4—Aerovox 1450
R_1—Ohmite "Brown Devil"
RFC—Bud CH-925

Figure 12, page 411
P.P. 224-Mc. Oscillator

C_1—Cornell-Dubilier 1-W
R_1—Ohmite "Brown Devil"
Tubes—Hytron

Figure 15, page 413
829 224-Mc. Transmitter

C_1—Solar MO
C_2—Solar MT
R_3—Ohmite "Brown Devil"
RFC_1—Ohmite Z-1
HY75—Hytron
829—RCA

Figure 18, page 414
56-Mc. Transmitter Exciter

C_1, C_2—Solar MP-4119
C_3—Cardwell ZU-75-AS
C_4—Solar MW-1239
C_5—Solar MW-1216
C_6—Solar MW-1210
C_7—Cardwell ZR-50-AS
C_8, C_{10}, C_{11}, C_{13}—Solar MW-1227
C_9—Solar MW-1233
C_{12}—Cardwell ZT-15-AS
R_1, R_4, R_7—IRC "BT"
R_2, R_5, R_6—Ohmite "Brown Devil"
RFC—Bud 920
Crystal—Bliley B-5
Tubes—RCA 6L6, Taylor T21

Figure 20, page 415
125-Watt Amplifier

C_1—Cardwell ET-30-ADI
C_2—Cardwell NP-35-ND
R_1—Ohmite "Brown Devil"
J_1, J_2—Yaxley 702

RFC—Johnson 760
T—Thordarson, T-19F98
Sockets—Hammarlund S-4
Tubes—Heintz & Kaufman
Bar knobs—Crowe

Figure 23, page 419
F.M. Exciter

C_1—Hammarlund MC-140-S
C_2—Hammarlund MC-50-S
C_3, C_8, C_{17}, C_{18}—Sprague 1FM-31
C_4, C_{12}, C_{14}—Sprague TC-11
C_5—Sprague TC-2
C_6, C_7, C_9, C_{10}, C_{11}—Sprague 1FM-25
C_{13}—Sprague UT-8
C_{15}—Sprague TA-25
C_{16}—Sprague TC-15
C_{19}—Sprague 1FM-35
C_{20}, C_{21}, C_{22}, C_{23}—Sprague TC-15
R_1, R_2, R_3, R_4, R_6, R_7, R_8, R_9, R_{11} to R_{18}, inclusive—Centralab 710
R_5—Centralab 516
R_{10}—Centralab 72-105
RFC_1, RFC_2, RFC_3—National R-100
IFT—Meissner 16-6211

Figure 26, page 422
Phase-Modulated Exciter

C_1, C_2—Johnson 75J12
C_3—Johnson 50J12
C_4—Hammarlund SM-100
C_5, C_9, C_{29}—Sprague TC-15
C_6, C_7, C_{11}, C_{12}, C_{15}, C_{16}, C_{17}, C_{20}, C_{21}, C_{22}, C_{23}, C_{25}, C_{28}—Sprague TC-11
C_8, C_{18}—Sprague 1FM-31
C_{10}—Hammarlund MEX
C_{13}, C_{27}—Sprague TC-1
C_{14}, C_{24}—Sprague 1FM-23
C_{19}—Sprague 1FM-35
C_{26}—Sprague TA-10
R_1 to R_6, inclusive, R_8 to R_{21}, inclusive, and R_{23} to R_{27}, inclusive—Centralab 710
R_7—Sprague 5K "Koolohm"
R_{22}—Centralab 72-105
X_1, X_2—Bliley BC-3

Figure 29, page 426
815 F.M. Transmitter

C_1, C_2—Bud LC-1682
C_3—Bud LC-1662
C_4—Bud LC-1661
C_5—Dismantled Bud NC-890
C_6, C_7—Sprague 1FM
C_8—Centralab 910-Z
C_9 to C_{18}—Sprague 1FM
C_{19}—Sprague SM33
C_{20}—Sprague UT-8
C_{21}—Sprague TC
C_{22}—Sprague 1FM
C_{23}—Sprague TA-10
R_1, R_2, R_3, R_5, R_6, R_{22}, R_{23}, R_{24}, R_{25}—Centralab 710
R_4, R_7, R_8, R_{16}, R_{17}, R_{18}, R_{19}, R_{20}—Centralab 714
R_9, R_{13}, R_{21}—Ohmite "Brown Devil"
R_{10}, R_{12}, R_{14}, R_{15}—Centralab 516
R_{11}—Mallory-Yaxley type N
RFC_1, RFC_2—Hammarlund CHX
RFC_3—Ohmite Z-1
T—Thordarson T-19F99
Meter Sw.—Mallory-Yaxley 151-L
Tubes—RCA throughout

CHAPTER 24
TEST and MEASURING EQUIPMENT

Figure 1, page 491
Ohmmeter

Resistors—Ohmite
Switch—Mallory-Yaxley 3100-J

Figure 8, page 494
112-Mc. Wavemeter

C_1—Bud type MC-926
C_2—Centralab type B, 3-12 $\mu\mu$fd.
Panel bearing—Bud type 532
Coupling—ARHCO type 250

Figure 11, page 497
Frequency Spotter

C_1, C_2—Hammarlund "Star" SM-100
C_4, C_5—Solar type MW
C_6, C_7—Solar type MP "Domino"
C_8, C_9, C_{10}—Solar type MW
C_{11}, C_{12}—Solar type MP "Domino"
C_{13}—Solar type MW
C_{14}, C_{15}—Solar D-820 electrolytic
R_5, R_6—Ohmite "Brown Devil"
S_1—Centralab 1465
L_1—Meissner 17-6753 b.f.o. coil
T_1—Thordarson T-13R11

Figure 13, page 498
Crystal Calibrator

X—Bliley SMC-100
L—Hammarlund CH-8
L_1—Hammarlund CH-X (altered)
C_1—Meissner 22-7002
C_2, C_3, C_4—Solar "Sealdtite"
C_6—Hammarlund SM-25
C_7, C_8—One Solar LGS-44
T—Thordarson T-13R01
Tubes—RCA

Figure 16, page 500
Field Strength Meter

Variable condenser—Hammarlund "Star"
Coil form—Bud type 906
Tube—RCA

Figure 20, page 502
Sensitive F.S. Meter

C_1—Hammarlund "Star"
C_2—Cornell-Dubilier 1-W
Coil form—Hammarlund XP-53

Figure 23, page 503
Grid Leak F.S. Meter

C_1—Cardwell ZR-50-AS
C_2—Sprague 45-35
C_3—Sprague 1FM-35
C_4—Sprague 1FM-21
R—Centralab 516
Cabinet—Bud 999 with chassis
Feed-through insulators—Johnson 44
Batteries—Burgess B-30, 5360, and "Little Six"

Figure 26, page 505
Frequency Meter, Monitor

C_1—Bud type MC-1852
C_2—Bud type MC-1876
C_3, C_4—Cornell-Dubilier type W
R_1—Centralab 710
R_2—Centralab 72-105
T—Thordarson T-13A34
S_1—Centralab 1405
Cabinet—Bud type 1746, with panel
Chassis—Bud CB-39
Batteries—Burgess type 4FL, 5360, and M30
Tuning dial—National type B
Tubes—RCA

Figure 31, page 508
'Phone Test Set

S_1—Yaxley selector type
C_1—Bud type 906
Dial—Crowe type 292
Tube—RCA

Figure 33, page 509
Keying Monitor

Speaker—Wright DeCoster N5LBU
C_1, C_2, C_3—Sprague TC
R—Centralab 710

Figure 35, page 510
R.F. and A.F. Power Meter

R—Ohmite D-100
M—Weston 425

Figure 37, page 511
Test Signal Generator

C_1—Meissner 21-5224
C_2—Meissner 21-5164 with 3 plates removed
C_3—Solar MOS-100
C_4, C_6—Solar S-0228
C_5—Solar type MW
C_7—Solar type M-240
R_1—Centralab 710
R_2—Centralab 516
S_1, S_2—Centralab 1462

Figure 40, page 512
Vacuum-Tube Voltmeter

C_1, C_4, C_5—Sprague TC-15
C_2—Sprague TC-1
C_3—Sprague TC-5
C_6—Sprague UT-8
R_1—Centralab 514
R_2, R_3, R_4, R_5—Centralab 710
R_6, R_7, R_8—Centralab 516
R_9—Centralab 72-107
R_{10}—Ohmite "Brown Devil"
T—Thordarson T-13R11
BC—Mallory
S_1—Centralab 1401
S_2—Bud SW-1115
Cabinet—Bud 993
Chassis—Bud CB-996
Tubes—RCA

Figure 48, page 517
Audio Oscillator

C_1—Radio Condenser Co., 4 gang
C_2—Solar MO-1406 and MO-1410 in parallel
C_3, C_5—CD or Solar S-0267
C_4, C_6, C_7, C_8, C_9—Solar DT-859A
C_{10}, C_{11}—Solar D-820 in parallel
½-watt resistors—Centralab 710
1-watt resistors—Centralab 714
10-watt resistors—Ohmite "Brown Devil"
R_{20}—Centralab 72-107
RL_1, RL_2—General Electric Mazda type S6
T_1—Thordarson T-13R11
T_2—Thordarson T-57S01
CH—Thordarson T-13C28
S_1, S_2—Yaxley 3226J
Cabinet—Bud C-1747
Chassis—Bud CB-997
Dial—National type ACN
Tubes—RCA

Figure 50, page 519
Cathode-Ray Modulation Checker

Resistors—Centralab 514, 516
T—Thordarson T-92R33
C_2—General Electric Pyranol type
R_2, R_3, R_7—Yaxley universal type
C_1—Solar "Domino"
Tubes—RCA

Figure 54, page 522
902 Oscilloscope with Sweep

R_1—Yaxley Y25MP
R_3, R_{22}—Yaxley Y50MP
R_7, R_8—Yaxley Y100MP
R_{20}—Yaxley UC506
R_{28}—Yaxley UC504
R_{31}—Yaxley Y500MP
All tubulars—Solar "Sealdtite"
Filter condensers—Solar DE908
S_2—Yaxley 3215J
S_4, S_5—Yaxley 3234J
T_1—Thordarson T-92R33

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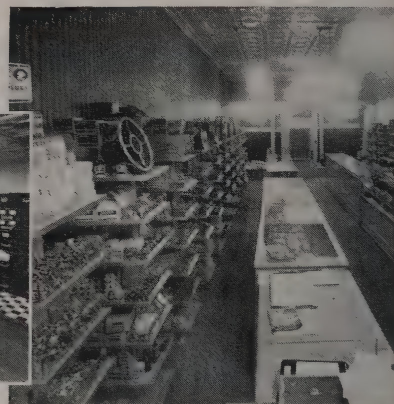

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